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CHAPTER 5

F-M RECEIVERS

Section I. INTRODUCTION

52. General

The operation and general circuitry of superheterodyne receivers for f-m communication form the subject matter of this chapter. The circuits used in the various operating parts of the superheterodyne receiver are discussed and comparisons are made between a-m and f-m circuits.

53. Superheterodyne

The separate stages required for the operation of an f-m superheterodyne are shown in the block diagram of figure 107. The r-f signal voltage coming from the antenna is amplified by the r-f amplifier stage and combined with the local oscillator voltage in the converter or mixer. The new frequency produced by this combination is the difference frequency or i-f (intermediate frequency) of the two signals, and is amplified by one or more i-f amplifiers. The greatly amplified i-f signal then is applied to an f-m detector which produces an audio-voltage output from the frequency variations of the i-f signal. This is applied to the audio amplifier, and the output of the amplifier is converted to the original sound by a loudspeaker or earphones.

54. Receiver Rating

a. Sensitivity and Selectivity. The sensitivity of a receiver is determined by the minimum signal voltage (in microvolts) at the input that is required to produce a specific output. The

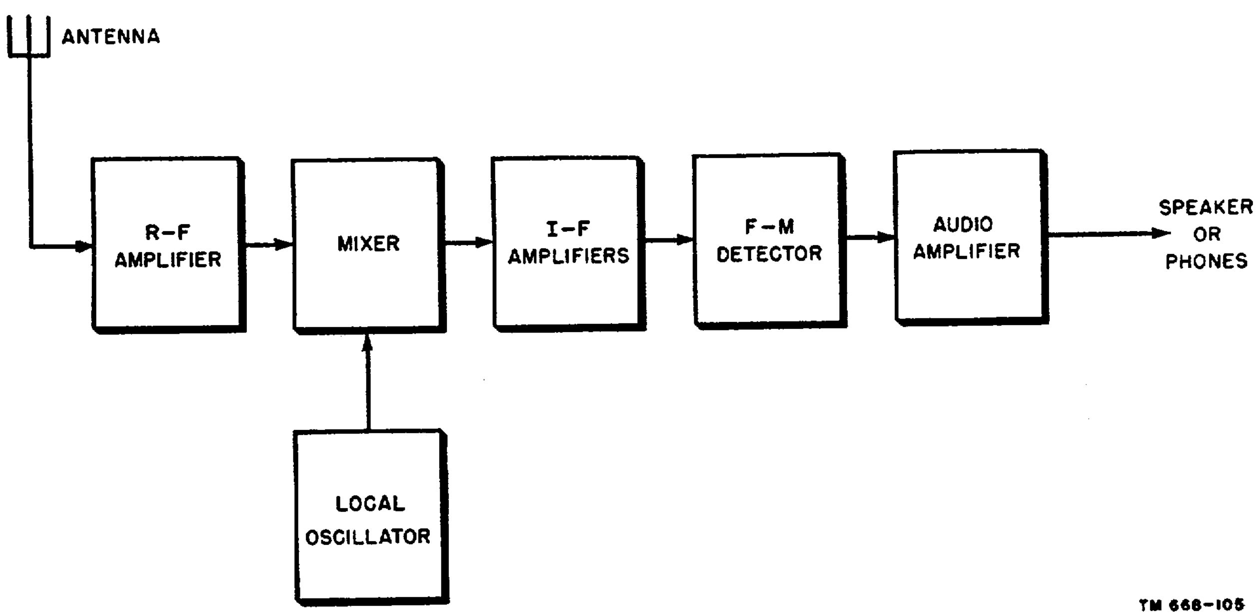


Figure 107. Block diagram of f-m receiver.

receiver producing the specified audio output with the smallest value of signal input is the most sensitive. The selectivity of a receiver is defined as the degree to which it is capable of discriminating against signals whose frequencies are other than that of the desired carrier. The selectivity for adjacent channels is determined largely by the characteristics of the i-f amplifier. Discrimination against image carriers depends on the circuit design of the r-f amplifier and the mixer, or converter.

b. A-M Response. Noise originating in actual defects, such as poor contacts or faulty parts, as well as the noise produced in the tubes, limits the maximum sensitivity of the receiver. Every f-m receiver, no matter how perfect, has some undesirable response to a-m signals, and these variations must be removed by limiting devices. The degree to which the receiver discriminates against amplitude variations in the received signal caused by fading or noise is a measure of the merit of the receiver.

55. Double Superheterodynes

Most f-m equipment operates in the v-h-f band, and at these frequencies it is not always possible to obtain optimum performance with the standard superheterodyne circuit. When good adjacent-channel selectivity is necessary, a low i-f is desirable; this, however, lowers the receiver rejection of image signals. Similarly, if good rejection of image frequencies is required, a high i-f should be used, but this is not compatible with good adjacent-channel rejection. These difficulties can be overcome by combining the advantages of high and low i-f amplification. The r-f signal is mixed with a local oscillator to produce a *high* i-f in the conventional manner. A *second* local oscillator voltage then mixes with the high i-f signal to produce a *low* i-f, as shown in figure 108. The double superheterodyne or double-conversion superheterodyne requires frequency conversion in two separate mixer or converter stages. To meet stringent military performance requirements, double superheterodynes commonly are used.

56. Auxiliary Circuits

Apart from the operational circuits, there are auxiliary circuits that provide automatic frequency control. To prevent operator fatigue, circuits are incorporated which silence, or squelch the noise that appears in the output of the receiver when no carrier is present. The special audio requirements of carrier telephone equipment also must be analyzed, since an important class of f-m equipment exists as a link between units using this type of telephone service.

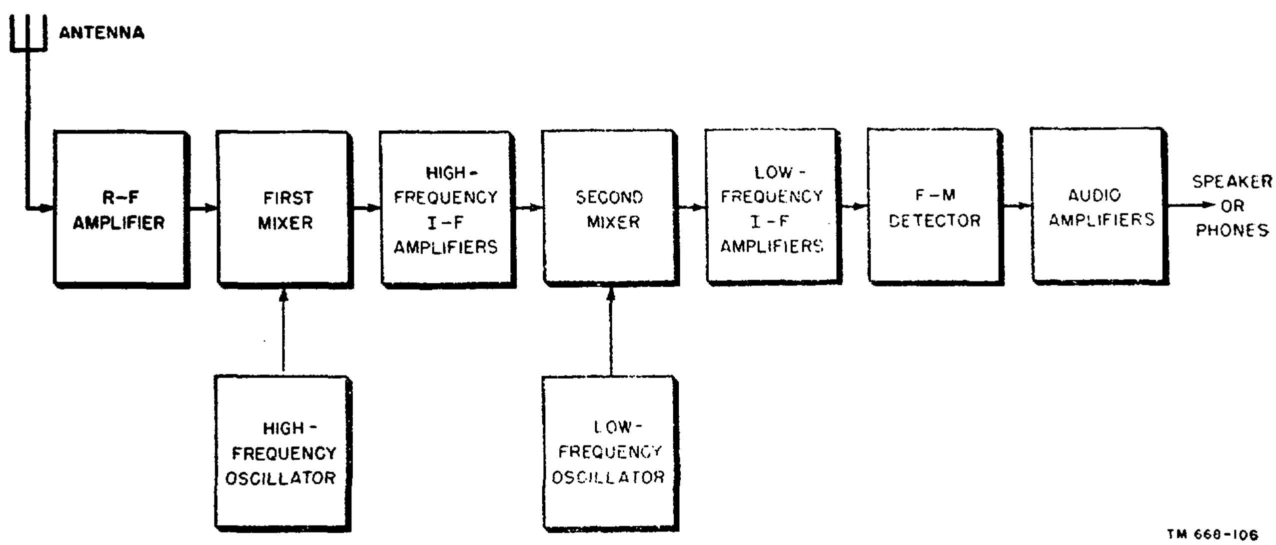


Figure 108. Block diagram of double-conversion f-m receiver.

Section II. R-F AMPLIFIERS

57. Basic Functions

a. The basic functions of the r-f amplifier are to increase the signal-to-noise ratio, to provide adequate image-frequency rejection, and to suppress local-oscillator radiations. These functions determine the design of the many different r-f circuits that have been developed, and each r-f amplifier exhibits varying degrees of performance as regards these factors. Considerations of tuning range complexity and stability also influence the choice of a particular circuit.

b. Since the r-f amplifier receives a signal at the lowest level of any stage in the receiver, any noise or other disturbance introduced in this stage has a proportionately greater effect. The performance of the receiver in respect to weak signals depends on the performance of the r-f amplifier, or the signal-to-noise ratio of its output. The r-f amplifier also must reject images and other unwanted frequencies that can reach the frequency converter and produce spurious responses. The efficiency of rejection depends largely on the design of the tuned circuits used. The local oscillator used in superheterodynes causes considerable interference if its signal is allowed to travel back to the antenna. Therefore, the r-f amplifier must be designed to block the signal from the local oscillator and prevent it from radiating. This is important when several receivers are operated close together on many different frequencies.

58. Sources of Noise

a. Fluctuation Noise.

(1) In any conductor, the movement of the electrons which constitute the current flow through the conductor is erratic and random. This random motion of electrons in a conductor or in the electron stream of a vacuum tube gives rise to small fluctuations in the voltage developed across the resistance of the conductor. The random variation of voltage is called fluctuation noise.

(2) No definite waveform or frequency can be associated with the fluctuation noise. However, the average amplitudes at all frequencies (noise spectrum) usually can be specified. Since the noise is distributed uniformly in respect to frequency, the amount of average noise voltage depends on the bandwidth. The fluctuation noise present determines the minimum signal to which a receiver will respond. The signal voltage must be sufficiently large to prevent the noise voltage from overriding the modulation on the carrier.

b. Thermal Noise.

(1) The molecules of any physical substance are always in violent motion, the average rate of this motion being perceived as temperature. Motion of the electrons caused by heat produces *thermal noise* across the terminals of any conductor containing resistance to the electron motion. From Ohm's law, the greater the value of the resistance in ohms, the more voltage developed across the resistor. Therefore, as the motion of the electrons in the conductor increases with temperature, the amplitude of the thermal noise in the output increases. The effect of the thermal noise, which is a type of fluctuation noise, depends on the bandwidth of the circuit which develops the noise.

(2) Figure 109 gives the amplitude of the average noise voltage usually encountered in the input circuits of f-m receivers. The temperature is assumed to be normal room temperature, or 68° Fahrenheit. The bandwidth of the input circuit can be estimated from a graph of its response characteristic. For example, it is common for an f-m receiver to have a bandwidth of 200 kc and an input circuit resistance of 300 ohms. As indicated in the chart, a noise voltage of about 1.2 microvolts results. This receiver cannot produce a useful output with incoming signals under 1.2 microvolts because of the threshold effect, and a higher value of input is needed.

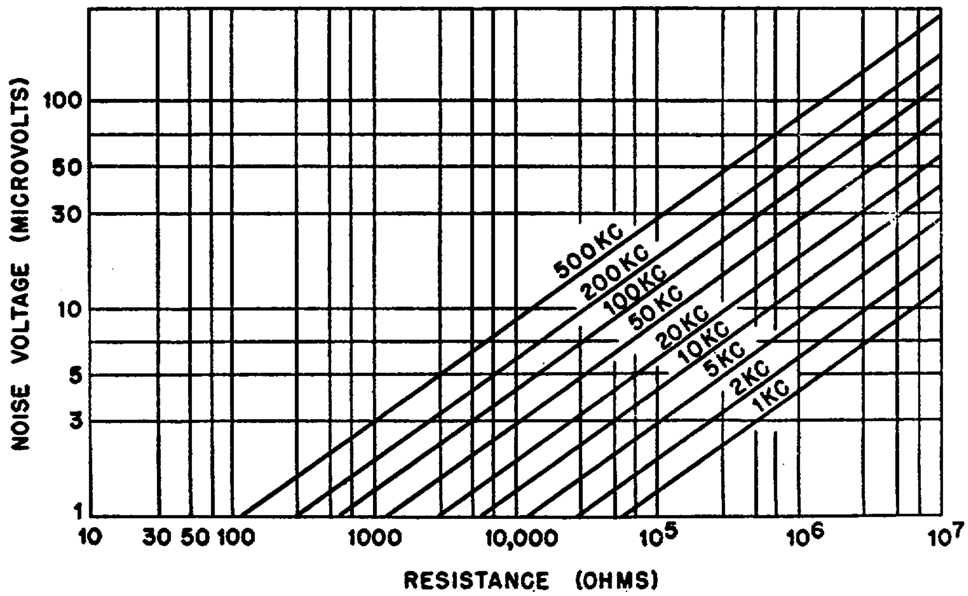


Figure 109. Noise voltage for different bandwidths and impedances.

c. Tube Noise. Fluctuation noise in vacuum tubes can be divided into three classes—shot noise, partition noise, and induced grid noise. Shot noise or shot effect occurs because the emission of electrons from the cathode is not uniform. Since the electrons leave the cathode at random instants of time and with random variations in velocity, their arrival at the plate is not uniform. Therefore, these variations in current flow represent a noise current and will produce a noise voltage when current flows through the load resistance. Partition noise is caused by the irregularities in the distribution of current in various positive electrodes in the tube. This noise is present only in pentodes, tetrodes, and other multigrid tubes. Variation in the number of electrons passing a control grid at high frequencies will induce large noise currents. This induced grid noise can be added to the plate current noise as though they were independent. The noise produced by a tube varies with changes in the potentials on its elements. A factor that strongly affects the amount of noise produced is the space charge in the tube. The noisiest condition is produced with temperature saturation; that is, when all of the available electrons are drawn to the plate and there is no space charge at all. It is the presence of space charge that makes amplification possible, and, fortunately, space charge also reduces the noise caused by random fluctuation in the electron stream.

d. Equivalent Noise Resistance.

(1) To permit direct comparison of the noise performance of standard tubes, a factor has been worked out that is known as the equivalent noise resistance of the tube. The value is that of a resistor that will produce the same amount of thermal agitation noise as that produced by the tube from all causes. The following chart indicates the range of values of this resistance for particular tube types.

Tubes	Ohms
Triodes	200 to 3,000
Sharp cut-off pentodes	700 to 7,000
Remote cut-off pentodes	2,400 to 14,000
Hexodes and heptodes	190,000 to 300,000

(2) The greater the number of positive grids, the greater the amount of noise produced by the tube, and the higher the equivalent noise resistance. The basic cause for this is the larger number of current divisions with increased partition noise. The noise produced by this equivalent resistance must be added to that introduced in the input

circuit of the tube to evaluate the performance of the r-f amplifier at low signal levels. In the previous example, the input circuit noise was 1.2 microvolts. Assume that this is applied to a pentode amplifier tube with an equivalent noise resistance of 3,000 ohms. From figure 109, the noise voltage corresponding to this resistance at the assumed bandwidth of 200 kc is 3.2 μv (microvolts). This is added to the input noise for an over-all noise of 4.4 μv.

59. Noise and R-F Amplifiers

a. General. Amplifiers do not discriminate between signal and noise within their bandwidth, but amplify them equally. For example, assume that the voltage amplification of the tube with its associated circuits in the previous example is 10 and that the tube delivers its output voltage into a similar amplifier. What will be the over-all noise performance of the system? With a noise voltage of 4.4 μv, a signal of 10 μv is applied at the input of the amplifier. Because the noise is of random phase it will be sometimes in phase with the applied signal and sometimes out of phase with it. Therefore, the signal in the output of the amplifier will be a combination of signal and noise which varies from the amplified sum to the amplified difference. The over-all output is the quadrature sum of the signal and the noise voltages, multiplied by the stage-amplification, or $\sqrt{10^2 + 4.4^2} \times 10 = 10.7 \times 10 = 107\ \mu v$. Since the second stage of amplification was assumed to be identical with the first, it adds 3.2 μv of noise to the applied signal of 107 μv. The output of the second stage is

$$(\sqrt{107^2 + 3.2^2})\ (10) = (107.1)\ (10) = 1070\ \mu v$$

The contribution to the total noise output of the 3.2 μv of noise added by the second stage is obviously very small. Therefore, the noise performance of the two stages together is essentially that of the first stage alone. If the second stage is the mixer in a superheterodyne circuit, it follows that the noise performance of the receiver is determined almost entirely by the first r-f amplifier. In general, if the r-f gain exceeds 5, the effect of the second stage is slight; if it exceeds 10, the effect is negligible.

b. Noise Figure.

(1) The noise figure expresses the relative merit of a receiver in comparison with a so-called perfect receiver. The perfect receiver is one which adds no noise to that produced by the antenna resistance and has a noise figure of 0 db. The quantity normally is expressed as a power ratio converted to decibels, and the smaller the noise figure, the better the receiver. Generally, the higher the frequency, the harder it is to obtain a good noise figure. Best attainable values range from 10 to 12 db at centimeter wavelengths, 6 db in the upper portion of the v-h-f range, where most f-m equipment is used, and below 3 db for frequencies under 30 mc. A perfect receiver has a noise figure of 0 db.

(2) The noise figure does not depend on the bandwidth of the receiver under test. A receiver that has a bandwidth of 200 kc has no greater sensitivity than one with a bandwidth of 15 kc if it has the same noise figure. However, because of the greater amount of noise contained in a wider bandwidth, it is harder in practice to obtain as low a noise figure for a wide-band receiver.

c. Tube Types and Noise Figures.

(1) The noise figure of an r-f amplifier is affected by many conflicting variables, and compromises must be reached. The tube is the ultimate factor which dedetermines the minimum noise figure that can be achieved, the transconductance and the input conductance being the principal factors in determining the noise figure. For good noise figure, it is desirable that the r-f amplifier have high gain so that the noise added by the following stages will be negligible. The higher the transconductance, the more gain from the stage. The input conductance of a tube is defined as the equivalent impedance that is seen looking into the grid. Effectively, it is composed of a resistance and a capacitance in parallel, placed across the tuned input circuit, and it

decreases as the operating frequency increases. Doubling the frequency decreases the input resistance by a factor of almost four; tripling it decreases the resistance by a factor of nine, and so on.

(2) It is common to use tuned transformers in the input circuits of r-f amplifiers to provide a voltage step-up from the low-impedance transmission line to the input impedance of the grid. Since the input impedance of the grid varies with frequency, the amount of voltage step-up that can be obtained decreases as the input conductance increases. Since no noise is generated in the input transformer, and the amount of noise generated by the tube is constant, the higher the input voltage applied to the tube, the lower the effect of tube noise, and the better the noise figure. It is desirable to use tubes that have a low input conductance. This permits a high voltage step-up in the input transformer with consequent overriding of tube noise. The input conductance will vary as the square of the frequency and has been found to depend on the input capacitance, the transconductance, and the inductance of the cathode load. For the lowest input conductance, all of these factors should be minimized. However, to obtain gain from the amplifier, there must be a reasonable amount of transconductance, and the choice of these values is necessarily a compromise. Since the noise produced within the tube is caused largely by irregularities in the electron stream, this stream must have great uniformity for low noise figures. The relative performance of various tube types in common use is listed in the table of equivalent noise resistances. Ratings in terms of input conductance generally are not available for all types.

60. R-F Amplifier Circuits

a. General. The r-f amplifiers of f-m receiving sets operate in the class A region of the operating curve and do not draw grid current. The several possible connections for a tube used as an r-f amplifier depend on which element is common to ground. For the triode, the possibilities are grounded cathode, which is the conventional connection, grounded grid, and grounded plate (cathode follower). The pentode and tetrode also follow these classifications, since the screens and other grids usually are at ground potential for radio frequencies.

b. Input Circuit Operation.

(1) The input circuit of most r-f amplifiers consists of an r-f transformer with an untuned primary that is connected to the antenna. The secondary of this transformer is considered to be a tuned series-resonant circuit that is adjusted to resonate with the frequency of the desired signal. Since the transmission line from the antenna is connected to the primary and generally has a low impedance (50 to 600 ohms), and the grid-to-cathode impedance of the tube is much higher, there is an impedance mismatch with serious loss of signal. The r-f transformer provides the necessary impedance match between the antenna and the grid of the tube. Actually, there is a slight mismatch for optimum noise figure.

(2) The transmission line used to transfer the signal from the antenna to the primary of the r-f transformer may be either coaxial-line or parallel-wire transmission line. When the coaxial line is used, the outer conductor is at the same potential as the circuit ground and an unbalanced input is required at the receiver. If parallel-wire transmission line is used, the voltages on each conductor are equal and opposite to each other and 90° out of phase with the ground potential, and a balanced input is required at the receiver. Both input circuits provide a certain amount of discrimination against noise and other spurious frequencies picked up on the transmission line. The outside of the coaxial line ean act as an antenna and

receive signals on the shield conductor. However, this outer conductor is effectively grounded at the input circuit and therefore produces no voltage across the grid-to-cathode circuit of the tube. Balanced input circuits suppress noise picked up simultaneously on both conductors of the line. The noise voltage is in phase on both sides of the line. This means that the voltage in respect to ground is 180° out of phase. The balanced input circuit responds only to currents that are out of phase on the line and 90° out of phase to ground; therefore, the noise voltage does not get to the grid of the tube.

(3) The output of the series-resonant circuit of the secondary decreases for frequencies on either side of resonance. Therefore, the over-all response of the amplifier falls off as the frequency increases or decreases in respect to the operating frequency. This produces the characteristic bell-shaped response of voltage versus frequency that is referred to as the selectivity curve. The ability to reject signals that are far removed from the operating frequency depends largely on the sharpness of this curve. In general, the sharpness of resonance of a tuned circuit depends on the *Q* of the inductor and the extent to which the input conductance of the r-f amplifier tube loads the input circuit. However, it usually is possible to manufacture for high-frequency work inductors whose *Q* is sufficiently great that the resistive component of the input conductance has a greater effect on the selectivity than does the resistive component of the coil impedance.

(4) Although it is possible to obtain the maximum amount of selectivity with a given tube, it often is necessary to compromise, since the receiver does not operate on a fixed frequency, but is tuned over a considerable range. This is accomplished by using a coil with more than the optimum amount of inductance. The various tuned circuits in the r-f mixer and oscillator stages must be correctly tuned for any operating frequency in the desired range. This means that the tuned circuits of the local oscillator and the r-f amplifier must *track* correctly over the entire range.

(5) For receivers that cover a limited band of frequencies, it often is desirable to avoid the complications of tracked circuits. This can be accomplished by keeping the r-f amplifier tuned to a fixed frequency and varying the frequency of the local oscillator. The selectivity of the input circuit is made low enough to cover the entire band in question and is called a *broad-band input circuit.* When the primary and the secondary of a transformer are moved close to each other, the coupling is increased beyond the critical value, and the response curve begins to broaden out. Instead of the familiar bell curve, a broad, flat-topped curve appears and the circuit is said to be overcoupled. As the coils are moved closer together, they become extremely overcoupled, the center of the curve begins to drop, and the two peaks appear still further removed from the center frequency. Such circuits are called overcoupled input circuits, and curves for values of coupling are shown in B of figure 110. A simple resonant circuit cannot provide both impedance transformation and wide bandwidth at the same time, and overcoupled input circuits with double and even triple tuning must be used. The circuit shown in A is overcoupled and a capacitor is placed in series with the primary and adjusted to resonance at the operating frequency. This provides the bandwidth necessary for f-m reception. More complicated versions, which produce three peaks in the response curve, can be obtained with three tuned circuits in the input. In f-m

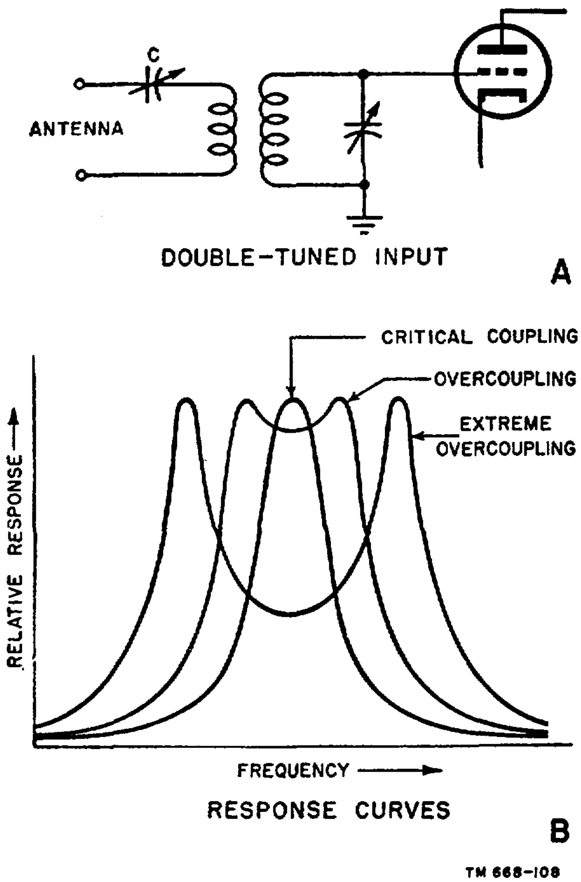


Figure 110. Coupled input circuit for r-f amplifiers.

equipment, the single- or double-tuned types are the most frequently encountered.

(6) At very-high frequencies, it is difficult to get the transformer type of input circuit to operate properly because of leakage and stray capacity. Therefore, the antenna or transmission line often is tapped directly on the coil. This connection (A of fig. 111) is suited especially to coaxial inputs. The operation is essentially that of an autotransformer, the coupling being between the two sections of the coil instead of through a separate primary as in a conventional transformer arrangement. Since the bottom of the coil is at ground potential and the top at grid potential, the input impedance increases as the tap is moved up the coil, and any value of line impedance can be matched. Usually, a small fixed capacitor is inserted in series with the tap lead to prevent stray d-c voltages that may appear on the antenna or transmission line from reaching the grid.

(7) If a large variety of antennas and transmission lines is to be used with the amplifier, it is necessary to provide an input circuit that can match a wide range of impedance. It is possible to construct a tuned transformer with a primary that can be moved in respect to the secondary. As the coupling varies, the effective input impedance also varies. Since this arrangement can be complicated mechanically, however, a variation of the tapped input transformer is used instead. To avoid the complication of a variable tap on the coil, two variable capacitors in series are used as tuning elements with the line connected to their junction, as in B. The capacitors act as a voltage divider. If the lower one has a much higher capacitance than the upper one, it has less reactance at the operating frequency. Therefore, the impedance across it is low and can be adjusted to match a wide variety of transmission lines. A variation of this circuit is shown in the pi-network of C, where the rotors of the two capacitors are grounded, with the input at the junction of the first capacitor and coil, and the output taken from the opposite junction. The pi-network passes all frequencies below the resonant frequency with only slight attenuation. This can be disadvantageous if the equipment is operated near powerful low-frequency transmitters. Input circuits for operation with balanced lines are shown in D, E, and F. Operation of the balanced circuits is similar to that of their unbalanced counterparts.

c. Grounded-Cathode Amplifier. The grounded-cathode r-f amplifier in A of figure 112, is the most familiar of all types. The input may be applied with any one of the many dif-

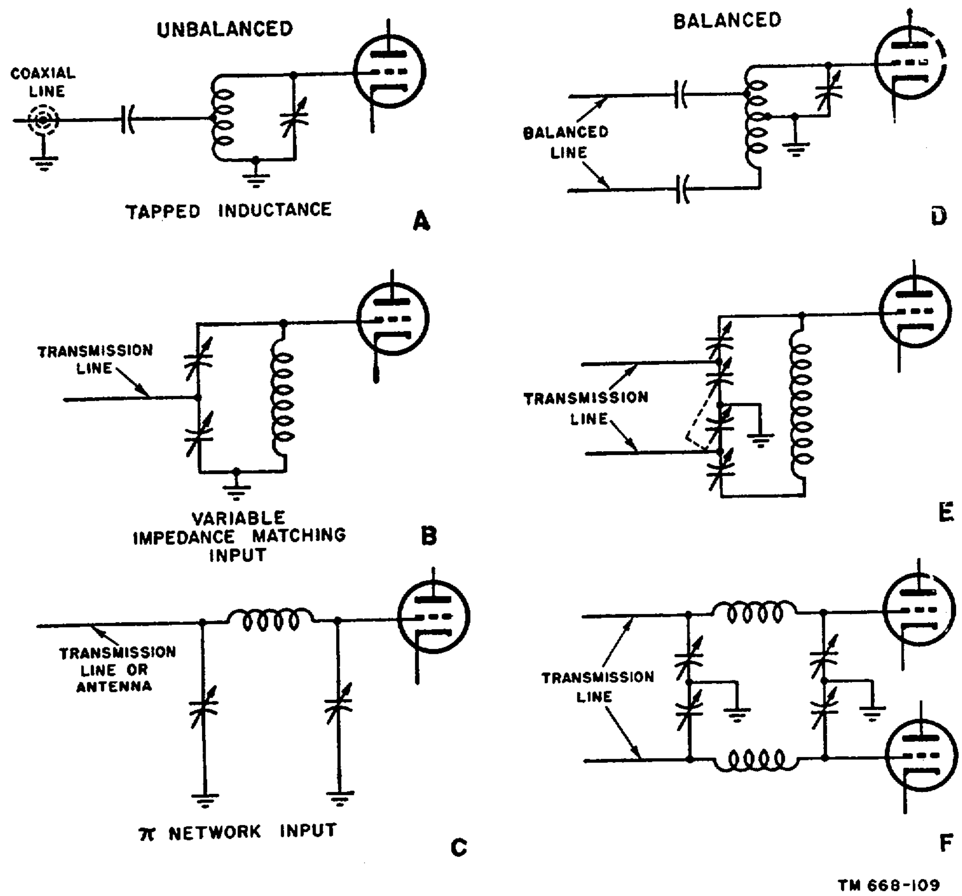


Figure 111. Miscellaneous input circuits for r-f amplifiers.

ferent input circuits to the grid, and the output is taken from a resonant load circuit in series with the plate. If a triode is used in this way, it must be neutralized to prevent tuned-plate tuned-grid oscillation. This neutralization seldom is successful over a wide frequency range, and the circuit generally is used only with pentodes. The noise performance of this circuit is good when triodes are used, but, with pentodes which have inherently higher noise figures, the arrangement is poor compared with that of other r-f amplifiers.

d. Grounded-Grid Amplifier. The grounded-grid circuit, shown in B of figure 112, permits the use of the triode with its lower noise figure, and does not require neutralization. However, the voltage gain of the amplifier is not as great as that of the grounded-cathode circuit because the input impedance is very low. The tuned circuit has little voltage step-up to overcome tube noise and the over-all noise performance suffers. The low-impedance input circuit permits the attainment of wide bandwidth and a reasonable noise figure without sacrificing too much voltage gain in the input circuit. The gain of the grounded-grid amplifier may not be great enough to override the noise produced by some converter tubes; therefore, it is common practice to find two grounded-grid r-f amplifiers in cascade. The added complications arising from this necessity and the need for special tubes limit its use. The tubes themselves must have very low effective plate-to-cathode capacitance if the shielding effect of the grounded grid is to be realized.

e. Cathode Follower. The amplifier shown in C has the input applied between the grid and ground, the plate is grounded for r-f, and the output circuit is in series with the cathode. The voltage gain of the stage is always less than unity, because the voltage variation at the cathode in an amplifier always is less than that

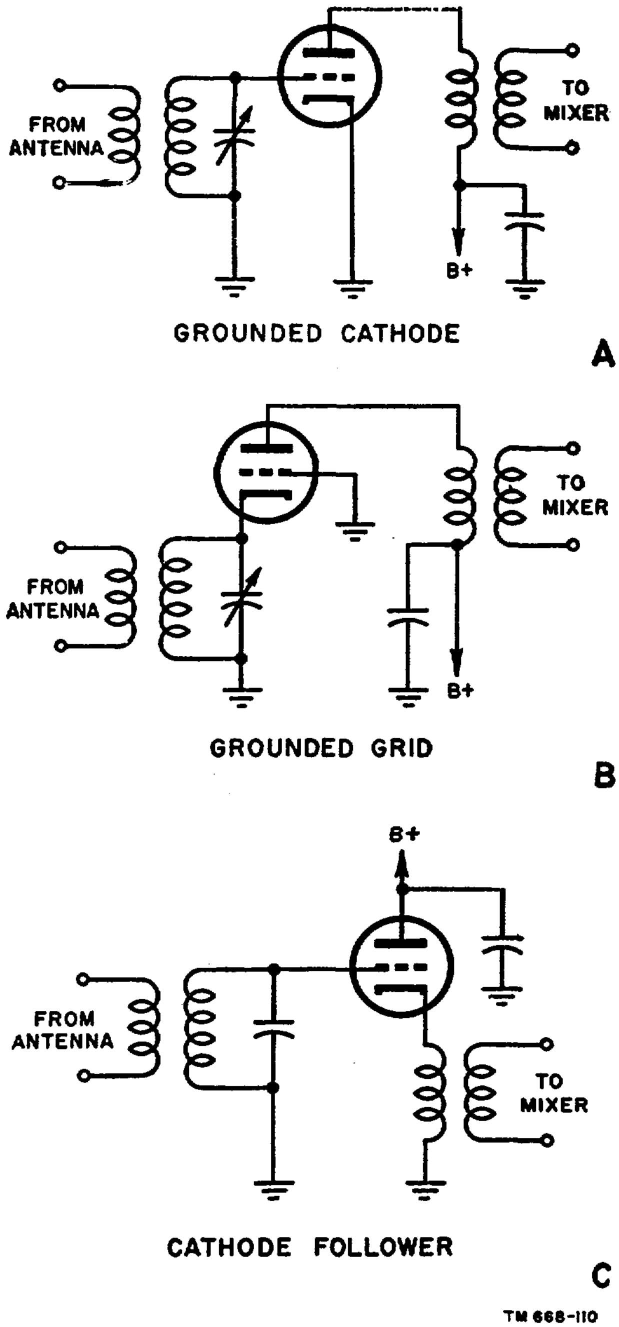


Figure 112. Single-tube r-f amplifier circuits.

at the grid. The output impedance of this amplifier is low because the cathode voltage is low and the current high. The input impedance is many times greater than the conventional grounded-cathode amplifier. Therefore, the input circuit can have a higher impedance and a higher voltage gain. The loss in the tube output, however, cancels out this advantage, giving a net result not very different from that of the conventional grounded-cathode amplifier. The cathode follower can be used with a triode without fear of oscillation. The low-impedance output has to be stepped up with an additional tuned transformer to match the input of the converter tube. Therefore, there is some danger of instability in coupling between the high-impedance transformers in the input and output circuits.

f. Single-Tube Amplifiers. Except for some differences in stability, input bandwith, and the like, there is little difference in the sensitivity that can be obtained with the three types of amplifiers. The noise figure of the tube is essentially independent of the manner of circuit connection. Where the high gain of a pentode stage is needed, the grounded-cathode circuit is used. Where a good noise figure and broad input bandwidth or a good match to the transmission line is desired, the grounded-grid circuit serves well. If high input selectivity is desired, the cathode follower, which loads the input circuit the least, can be chosen.

61. Two-Tube R-F Amplifier Circuits

a. General. The grounded-cathode, grounded-grid, and cathode-follower amplifiers may be used together or in push-pull. The push-pull amplifier permits a low-impedance, balanced-input circuit and can be used at very-high frequencies. Circuit diagrams are shown in figure 113. When cascade amplifiers are used, it is possible to have nine different circuit arrangements, since the output of any one of the three amplifiers may be connected in three different ways. The circuits most frequently used are the double grounded-grid, the cathode follower into grounded-grid, the grounded-cathode into grounded-grid, and the double grounded-cathode circuits.

b. Push-Pull Amplifiers.

(1) The push-pull circuit has a balanced input and output, and therefore is used widely with a balanced line. Since input voltage for a push-pull circuit must be twice that for an unbalanced amplifier, and since the input impedance is four times that for a conventional circuit, the over-all re-

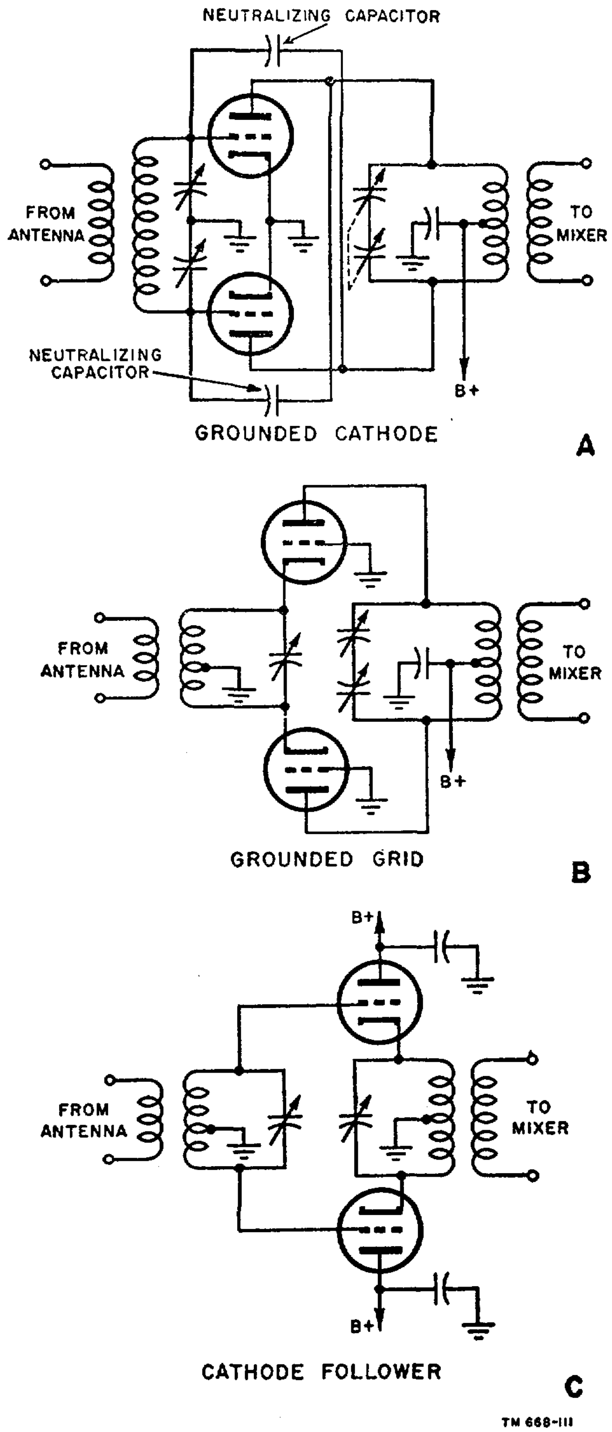


Figure 113. Push-pull r-f amplifiers.

quirement of impedance and voltage results in an input circuit with a higher ratio of inductance to capacitance. This becomes significant at very-high frequencies, where the internal tube capacitance can be an appreciable part of the circuit capacitance. The tube capacitance is halved effectively in the push-pull circuit, because the grid-cathode capacitance of the two tubes is in series across the input circuit. In the grounded-cathode circuit using triodes, it is necessary to use a cross-neutralization circuit. The over-all result is a circuit with a voltage gain that is much the same as that for a single neutralized triode, for the input voltage must be doubled even though the outputs of the two tubes add to each other.

(2) This implies that the noise produced by each tube also adds in the output. However, the voltage step-up in the input circuit is greater, and the over-all noise figure for the push-pull grounded-cathode circuit therefore is the same as that for each single-ended tube. The major advantage of the push-pull circuit, apart from its balance, is comparative ease of neutralization. Tubes such as the JAN types 6J6 and 12AT7 are specifically designed for these circuits.

(3) The push-pull grounded-grid amplifier, in B, permits the use of a balanced input circuit of very low impedance. This means that a broad bandwidth can be obtained easily. The voltage gain of the amplifier is the same as that for a single grounded-grid stage, and the noise figure is unchanged. This circuit can be used at very-high frequencies, where neutralization required for the grounded-cathode push-pull stage becomes too critical. The tubes used must have low plate-to-cathode capacitance, or the shielding effect of the grounded grid is lost. Moreover, where two triodes are incorporated in a single envelope, the cathode leads for each half must be brought out separately. JAN tube types 12AT7 and 6BQ7 sometimes are used in this way.

(4) Push-pull cathode followers are uncommon as r-f amplifiers because the

low-impedance output is not suitable for most mixer and converter tubes. However, when the mixer is a germanium or silicon crystal, it has a low-impedance input and the push-pull cathode follower can be used.

c. Two-Tube Cascade Amplifiers.

(1) Because of the deficiency in gain of a single grounded-grid stage, two stages generally are used in cascade. The output of the first is applied to the input of the second, as shown in A of figure 114. The over-all gain is the product of the individual stage gains. Because the gain of the first stage is seldom sufficient to permit it to overcome its own internally generated noise, the dual-cascade grounded-grid amplifier is not capable of as low a noise figure as a single neutralized triode. The arrangement does not require neutralization, and, because of the low impedance of the input circuits, it has a large usable bandwidth.

(2) The cathode follower driving a grounded-grid amplifier is referred to frequently as a cathode-coupled amplifier. In this connection, the tubes have a common cathode connection through the cathode resistor, as shown in B. The noise figure for this amplifier is good, although each stage contributes to the noise without enough added amplification to overcome it. Consequently, where low noise is necessary, this circuit is less desirable. The input impedance is high and the output impedance also is comparatively large. Only two tuned circuits are necessary, with the common cathode resistor acting as the interstage coupling. In addition, amplifier operation is stable without the necessity for neutralization. Where simplicity of circuit connection is required, the cathode-coupled amplifier combines fair performance with a minimum number of parts.

(3) The most important of the cascade amplifiers consists of a grounded-cathode triode driving a grounded-grid triode. This circuit usually is referred to as a driven grounded-grid circuit. From the standpoint of noise figure, it has the most favorable combination of characteristics. Several possible arrangements, shown in C, D, and E, permit this circuit to operate with stability. The amplifier is stable because of the low plate load of the input amplifier formed by the input of the grounded-grid second stage. The voltage gain of the first tube is negligible, and gain is obtained in the second stage with the noise figure of the first stage. If the amplification factor of each tube is large, the noise figure is that of the input triode alone. This circuit has all of the advantages of the triode grounded-cathode amplifier without any of the disadvantages, such as need of precise neutralization. However, the noise figure of the first stage can be improved further at higher frequencies if it is neutralized.

(4) The three diagrams, C, D, and E, indicate different means for accomplishing neutralization in the driven grounded-grid circuit. C shows an inductor and blocking capacitor connected in series between the grid and the plate of the first tube. The value of inductance is chosen to produce parallel resonance with the grid-plate capacitance at the operating frequency. A variant of this method of neutralization is shown in D. Here, an out-of-phase voltage from the output circuit is applied to the input circuit through a capacitor. This arrangement operates over a wide range of frequencies if the leads of the neutralizing capacitor are short. The circuit of E is similar to that of D except that the two sections of the tube are coupled directly from plate to cathode, and the plate current of the first tube flows through the second tube. This arrangement is the ultimate in simplicity and noise performance, since it uses very few parts. The

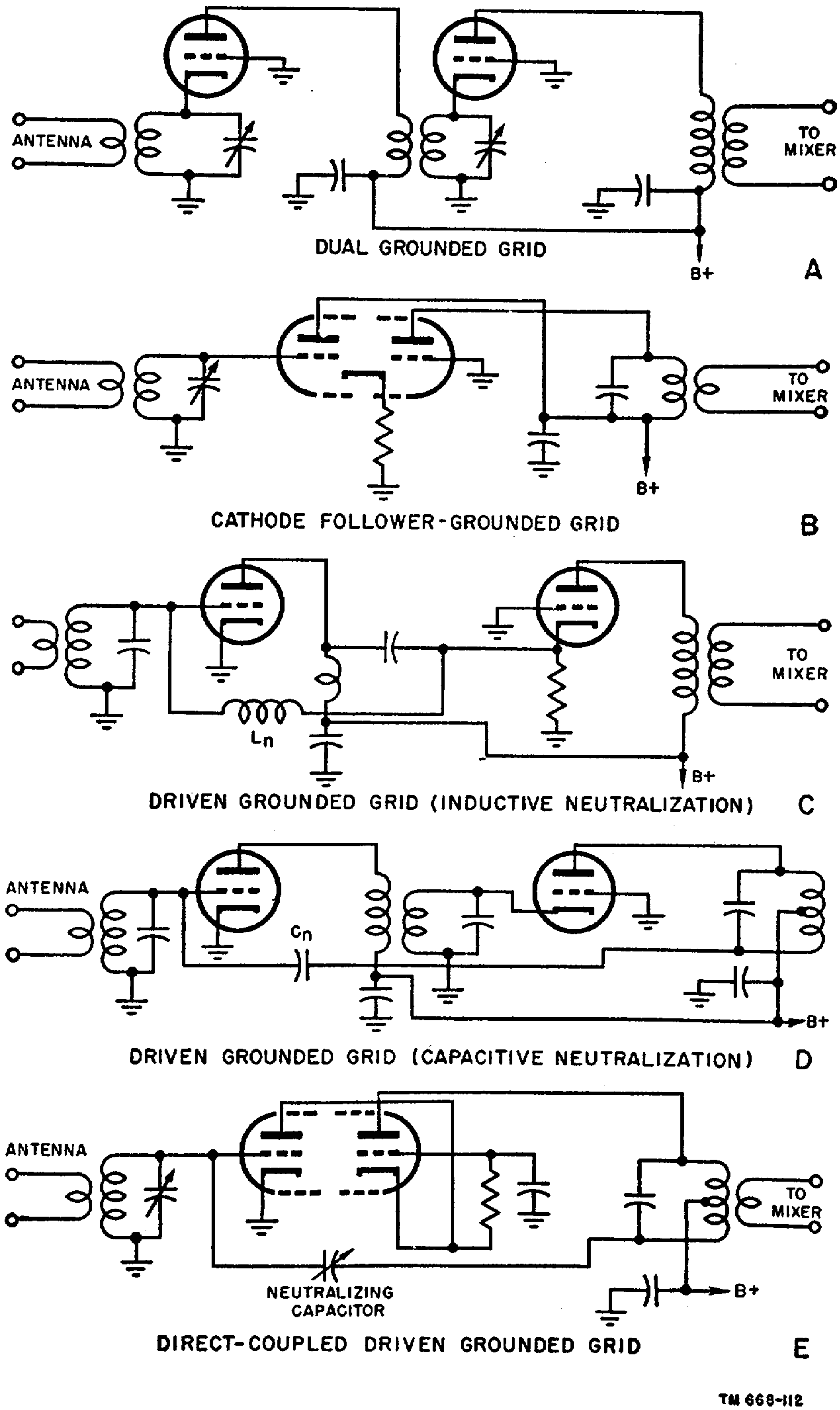


Figure 114. Cascade r-f amplifiers.

direct-coupled circuit combines the cut-off characteristics of both tubes. Therefore, variations in d-c grid voltage required to cut off the two tubes are greater than for a single tube alone. This reduces the susceptibility of the stage to overload from strong signals and reduces modulation between signals of different frequencies.

d. Automatic Volume Control for R-F Amplifiers.

(1) Automatic control of the gain in the first r-f amplifier of an f-m receiver is desirable because it tends to equalize the signal applied to the mixer or converter stage. In a-m receivers, avc (automatic volume control) is used to hold the over-all gain of the receiver to a constant level and to minimize fading and similar effects. In an f-m receiver, this is not necessary, because the f-m detector does not respond readily to fading in a signal. Therefore, the requirements placed on the avc circuit in an f-m receiver are not severe, and many f-m receivers do not contain avc.

(2) Since strong signals tend to overload the mixer stage, the f-m avc circuit applies a negative potential to the control grid of the first r-f amplifier. This negative potential increases with an increase in signal, reducing the gain of the stage. Although tubes with a remote cut-off characteristic give the best performance, they have poor noise characteristics. The direct-coupled dual-triode arrangement is a practical solution to this problem. Sharp cut-off tubes can be used, although smaller avc voltages must be applied to prevent cutting off the tube entirely on strong signals.

Section III. MIXERS AND CONVERTERS

62. General

a. Principles and Purposes of Frequency Conversion.

(1) The frequency converter in a superheterodyne receiver beats the signal from the r-f amplifier against the signal from the local oscillator to produce a signal at the intermediate frequency. Frequency conversion circuits are made up of two parts, the local oscillator and the mixer. The local-oscillator signal and the r-f signal amplitude-modulate the electron stream in the mixer. This action produces side bands equal to the sum and difference frequencies of the r-f and local-oscillator signals. The lower side band, which is the difference frequency, generally is used as the i-f. The local oscillator always is much stronger than the signal so that the percentage of modulation is low. This prevents the development of spurious side bands. If the oscillator signal contains harmonics, side bands also are produced between these and the signal. It is the function of the tuned input circuit to pass only the signal frequency, and the output circuit discriminates against all but the desired side band.

(2) Although separate local oscillators are used almost always at the higher frequencies, the functions of the mixer and the oscillator can be combined in one tube envelope. This tube is called a *converter* and usually is limited to intermediate frequencies below 15 mc in f-m receivers. In a double-conversion f-m receiver, however, the second mixer operates in the frequency range where a converter is practical.

b. Operation of Modulation Process.

(1) The low percentage of modulation required for frequency conversion can be produced in several ways. The method most frequently used depends on the *transfer* characteristic of a

tube or other circuit element. The transfer characteristic expresses the relationship between the signal ap-

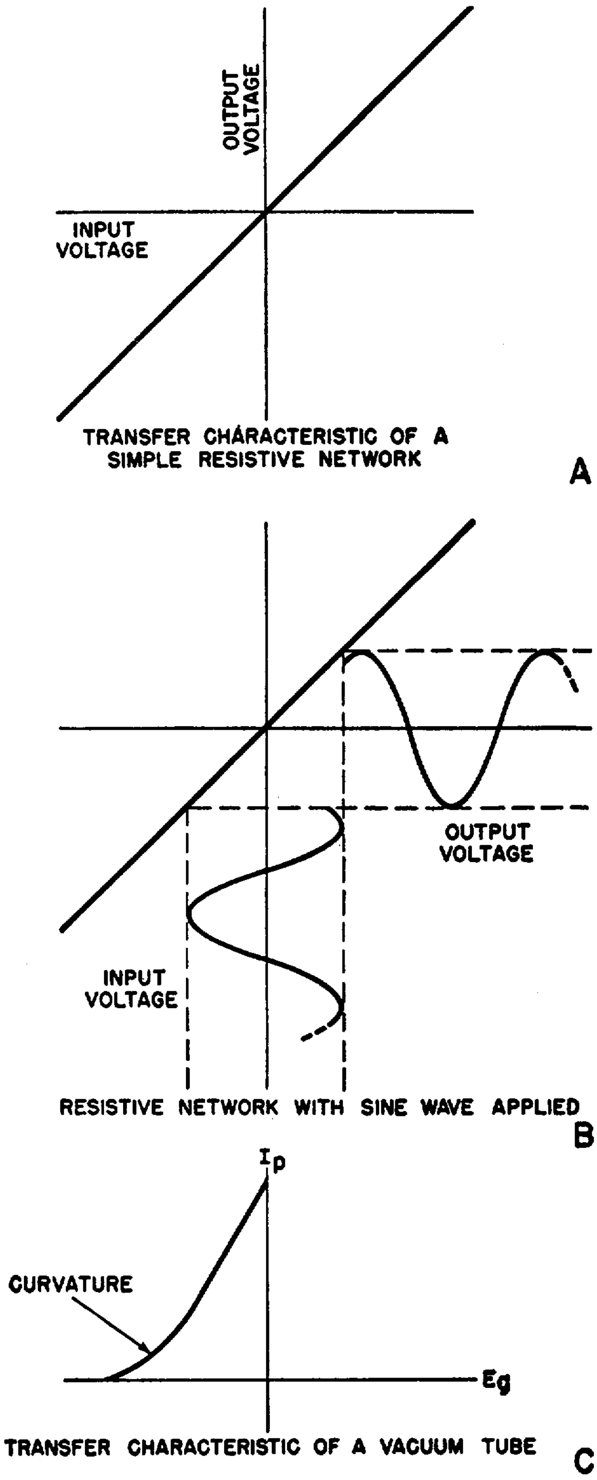


Figure 115. Characteristic response of resistive network.

plied to the input of a device and the signal obtained from its output. For example, a straight-line transfer characteristic of a resistive network is shown in A of figure 115. This is called a linear transfer characteristic and shows that the output voltage is directly proportional to the input voltage. A sine-wave input to this circuit results in the output shown in B.

(2) The transfer characteristic of a vacuum tube is not a straight line, since the relationship of e_g to i_p usually is curved at low values of plate current (C of fig. 115). Therefore, the vacuum tube is a *nonlinear* device. When the voltage on the grid of a vacuum tube becomes more negative and reaches the cut-off value, no current flows in the plate circuit. Consequently, for an entire range of voltages no current flows in the output circuit. Therefore, unlike the resistor, where current flows in proportion to any value of applied voltage, the vacuum tube is nonlinear, even if its transfer characteristic is perfectly straight. Other devices besides vacuum tubes are nonlinear, including silicon and germanium crystals.

(3) Any nonlinear device can be used as a mixer. In figure 116, two sine waves of different frequencies are applied to a nonlinear circuit. The output waveform is the familiar modulation envelope, which is equivalent to the carrier plus the side bands.

(4) There are two methods of producing the modulation envelope in a vacuum-tube mixer, and both utilize the electron stream. One method injects the r-f signal at the grid adjacent to the cathode and the local-oscillator signal at the same, or a following grid. The other method connects the oscillator to the grid adjacent to the cathode or to the cathode itself, and applies the r-f signal to an outer grid. Each method produces a variation of the electron stream by the oscillator, and a further variation by the r-f signal.

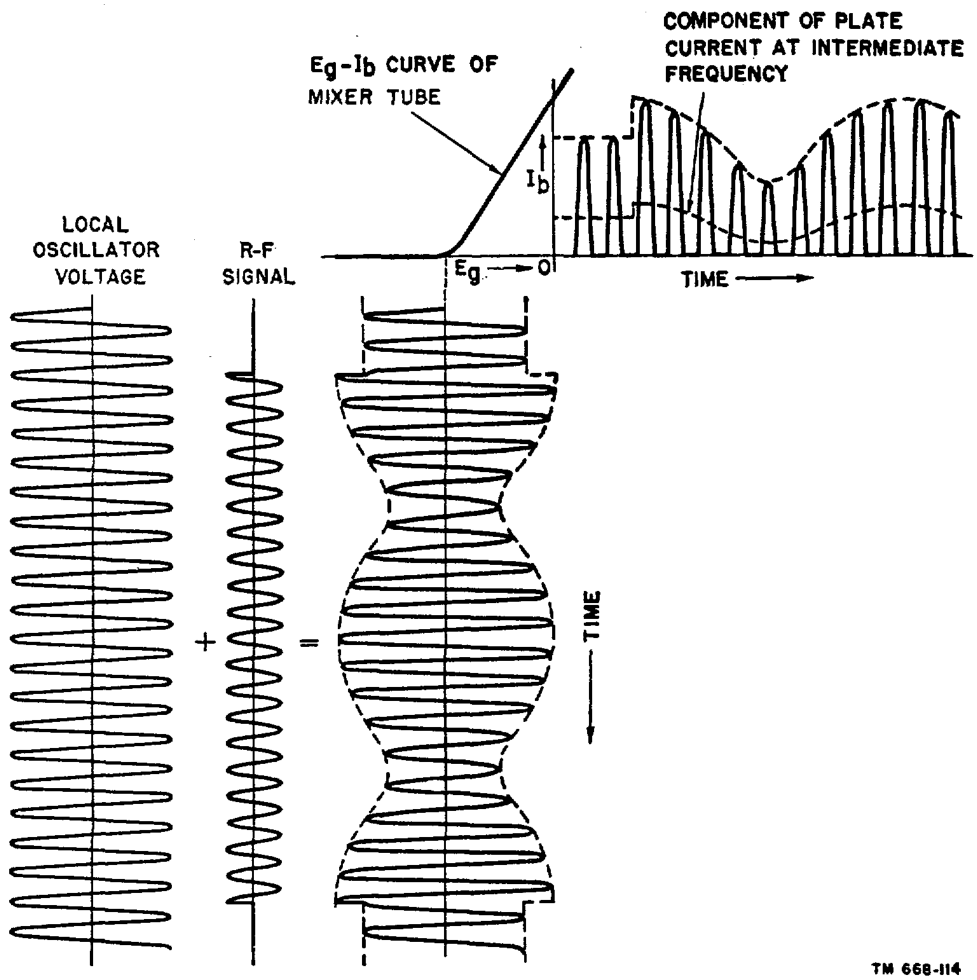


Figure 116. Mixer operation.

The electrons which reach the plate produce a current which has the shape of the modulation envelope. Effectively, these two methods do not differ, because of the nonlinear properties of an electron stream attracted to a plate.

c. Classification of Mixers. Mixers are classified according to the way in which the r-f and local-oscillator signals are applied. When both are applied to a single terminal, the mixer is called single-ended. If the r-f signal is applied to the control grid and the local-oscillator signal to an outer grid, the mixer is called *inner-grid* modulated. When the oscillator signal is injected before the r-f signal, the tube is said to be *outer-grid* modulated. It is desirable to inject the local-oscillator signal at a low-impedance point such as the cathode of the tube. This type of oscillator-signal connection is called a *cathode-injection* mixer.

d. Conversion Transconductance.

(1) The *conversion transconductance* of a mixer is defined as the ratio of the output current at the intermediate frequency, I_{if}, to the input signal, E_{rf}, at radio frequency with zero plate load impedance:

$$G_c = \frac{I_{if}}{E_{rf}}$$

Since frequency conversion is involved in a mixer, this must be indicated in the conversion transconductance. Strictly speaking, conversion transconductance is defined with small load resistance compared to plate resistance. As long as this condition is met,

changes in output are caused only by changes in load resistance. Maximum output at the intermediate frequency is obtained with cut-off bias on the control grid. Therefore, the operation of the mixer is similar to that of the plate detector. There is a very close similarity between conversion transconductance and the ordinary transconductance of an amplifier tube.

(2) The amplitude of the local-oscillator signal must be great enough to reach the part of the transfer characteristic with the steepest slope. As grid current begins to flow in the local-oscillator circuit, the input circuit of the mixer becomes loaded and therefore the local oscillator plus the r-f signal amplitude must be less than this amount. The mixer control grid must never be driven positive in respect to the cathode by local oscillator voltage. Bias for the mixer is obtained with a cathode resistor or, more frequently, with a grid leak. When a grid leak is used, the amplitude of the local-oscillator signal determines the d-c bias. The value of the a-c plate current which flows at the intermediate frequency, and hence the total output, depend chiefly on the slope of the transfer characteristic. The maximum conversion transconductance that can be reached is approximately 28 percent of the transconductance of the tube operating as an amplifier.

e. Conversion Gain.

(1) The output current of the mixer, as shown in figure 116, contains a strong component at the intermediate frequency. The tuned circuits between the output of the mixer and the i-f amplifier are parallel-resonant at the i-f frequency, and have a high impedance at the intermediate frequency and low impedance at all others. Therefore, the output current develops a large i-f voltage drop across this circuit. The conversion gain of the mixer is defined as the ratio of this i-f output voltage to the r-f signal input voltage:

$$A = \frac{E_{if}}{E_{rf}}$$

The amount of gain depends on the impedance of the tuned circuit. Therefore, it cannot be specified with a single constant for each tube type. A graph can be drawn, however, which shows this gain for different values of load impedance.

(2) The gain of the mixer depends on the conversion transconductance of the tube, since this determines the amount of current which flows through the tuned load circuit and the voltage developed across it for a given value of grid voltage. For high sensitivity, the mixer must have a high value of conversion transconductance. Similarly, a high value of load impedance also is desirable. The flow of current in the output circuit of the mixer is limited by the *a-c plate resistance* of the tube. This is the ratio of the a-c plate voltage to the a-c plate current at the intermediate frequency and is essentially in series with the tuned output circuit. When its value is more than five times that of the impedance of the output circuit itself, the current which flows is limited by the plate resistance alone. Therefore, the conversion gain does not depend on the plate resistance of the tube for low values of load impedance. As the load impedance is made larger, however, it begins to approach the value of the plate resistance, and the current flowing in the circuit decreases, with a corresponding decrease in the conversion transconductance. The load impedances which give the limiting value of gain usually are much higher than can be obtained with practical tuned circuits. Therefore, the maximum gain of the mixer stage as a whole is effectively determined by the maximum attainable impedance of the tuned plate load circuit.

63. Requirements for Mixers and Converters

a. Spurious Responses.

(1) Since a mixer is nothing more than a low-level modulator, the tuned input circuit can combine many signals with the fundamental of the local oscillator, or any of its harmonics, to produce an i-f output. If all but the desired signal are greatly attenuated before reaching the mixer input, and the oscillator is operating with low harmonic content, the response to spurious signals is minimized. This response depends on the selectivity of the input circuit of the mixer and of the preceding r-f stage. The over-all selectivity of the two stages is the product of the individual selectivities. If the response at a frequency far away from resonance is specified in decibels below that at resonance, the response of the two stages is the sum of the individual responses in decibels.

(2) The most important spurious response is the *image* frequency which combines with the local oscillator to produce a spurious i-f side band. If the local oscillator is higher in frequency than the desired channel by 1 mc, for example, the image response frequency is 1 mc higher, or 2 mc above the desired channel. If a strong carrier appears at the image frequency, it interferes with the desired station. The higher the intermediate frequency, the farther away the image is from the operating frequency and the greater its attenuation in the tuned circuits ahead of the mixer.

(3) Spurious responses of less importance than the image frequency can occur at many different frequencies where harmonics of the local oscillator beat with undesired signals (and their harmonics) to produce the i-f signals. Many other possible spurious responses such as these are troublesome when the receiver is operated in the vicinity of a strong transmitter. It is difficult in some instances to determine the cause of a particular response in a receiver.

b. Interaction of Oscillator and Signal Frequencies. In all mixers, a certain amount of coupling exists between circuits which introduce the r-f signal to the tube and those which introduce the local-oscillator signal. When the i-f is low compared with the operating frequency, the frequency of the oscillator and that of the mixer input circuit are very close together. If a strong signal appears in the mixer input circuit while the receiver is tuned to a weak signal on an adjacent channel, the stronger one tends to cause the oscillator to shift frequency and *lock in* with it. This results in failure of reception of the weak signal, and is called *pulling*. The degree of oscillator-frequency pulling is dependent on the i-f, on the coupling between the oscillator and input circuits, and on the basic stability of the oscillator itself. The condition is aggravated with avc applied to the mixer. In general, coupling between the oscillator and input circuits is greater at high frequencies, where oscillators tend to be less stable. Mixers have varying degrees of isolation between the oscillator and the r-f circuits. Therefore, the oscillator-mixer combination designed for use at high frequencies is strongly influenced by the degree of isolation. Converters, because of the association of oscillator and mixer in one tube envelope, generally are the worst offenders in regard to pulling.

c. Noise and Input Loading in Converters and Mixers.

(1) Like r-f amplifiers, the control grids of both mixer and converter tubes present a conductance which appears across the tuned input circuit. The amount of loading depends on the type of converter and the operating frequency. Where the r-f signal is introduced on an inner grid, with the local-oscillator signal on an outer grid, the loading is negative. This means that the resistive component of the conductance has a negative value at the operating frequency. The negative resistance tends to cancel out

some of the positive resistance of the tuned circuit. Therefore, its *Q* is raised, resulting in an improvement in image rejection. When the *Q* of a parallel resonant circuit is raised, its effective load impedance increases, and the output voltage from the r-f amplifier goes up. This improves the operation at high frequencies. In those mixers where the local-oscillator signal is injected on an inner grid, or on the same grid as the r-f signal, the input loading is positive. Here the action on the gain and image rejection is the reverse: Image rejection is lowered because of the lower *Q*, and the r-f gain is diminished because of decreased plate load impedance.

(2) Like r-f amplifier tubes, converters produce shot effect and partition noise. Because many converters have a larger number of grids than r-f amplifiers, the noise figure is higher. Even when triode tubes are used as mixers the noise figure is much lower than for the same tube operating as an r-f amplifier. This additional noise is contributed by the local oscillator. Any irregularity in the frequency of oscillation produces an effective noise voltage, in addition to the thermal, partition, and shot-effect noises generated in the oscillator tube and circuit. Generally, in f-m equipment for the v-h-f band, sufficient r-f gain is used ahead of the converter to prevent this noise from being a serious problem. Above the v-h-f range, however, r-f amplifiers are no longer practical, and crystal mixers which have a certain amount of loss but contribute nothing to the noise level are used. Of course, the local oscillator still will add noise, so that it becomes the main source of additional noise.

(3) The noise produced in multigrid mixers and converters can be reduced to a low value, equivalent in some cases to that of a pentode amplifier, by the proper application of feedback at the operating frequency. This feedback usually is generated across a tuned parallel-resonant circuit in the cathode lead. It is adjusted to resonance below the operating frequency so that its reactance is primarily capacitive. Variations in cathode current then appear as voltage variations across this tuned circuit. These voltage variations aid the voltage at the grid, effectively increasing the signal without adding to the noise. Therefore, the over-all noise figure is reduced. To prevent instability and oscillation, some of the signal frequency output from the plate circuit is returned to the grid through a neutralizing circuit. This does not interfere with the feedback that reduces noise. Similar arrangements can be used with crystal mixers to cancel out the noise introduced by the oscillator.

d. Oscillator Radiation.

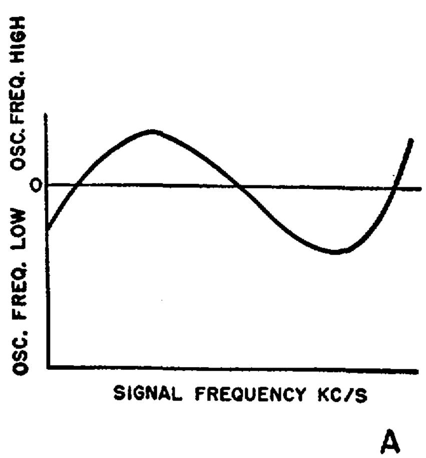
(1) The local oscillator in an f-m receiver generally operates at a fairly low level, producing only microwatts of power. However, even this low power, if allowed to reach the antenna, can cause serious interference to receivers operating in the vicinity. The position of a receiver with a radiating oscillator can be located by unfriendly forces through triangulation. Therefore, the suppression of oscillator radiation is important in military receivers. The mixer is most critical in this respect. Oscillator power can reach the antenna back along the path through which the received signal comes. The coupling of the oscillator and signal sections of the mixer influences this strongly. Where the oscillator voltage is injected on the same grid as the signal voltage, the condition is especially acute. In those mixers where the oscillator is applied to a separate grid, the coupling depends on the tube capacitance. Part of the oscillator voltage reaches the signal-tuned circuit through direct or stray-capacitive coupling. Another portion reaches

it through the electron stream of the tube.

(2) The suppression of oscillator radiation for a particular tube takes place primarily in the tuned circuits of the mixer input and the r-f amplifier. At low frequencies, where outside noise levels are far in excess of internal tube noise, the r-f amplifier is used to attenuate the local-oscillator signal applied to the antenna. Since the selectivity is greatest far from resonance in a tuned circuit, higher intermediate frequencies result in better local-oscillator suppression. However, where crystal mixers are used without r-f stages, as in u-h-f (ultrahigh-frequency) operation, the radiation must be eliminated entirely in the input circuit. The very-high intermediate frequency required to allow for lowered selectivity makes the design and operation of i-f amplifiers difficult. In general, doubling the intermediate frequency reduces oscillator radiation by a factor of 4. In vacuum-tube mixers, if the i-f is close to the operating frequency, the load presented to the oscillator voltage through the capacitance of the tube is high, and the radiated voltage increases considerably.

e. Tracking of Mixer and Oscillator Circuits.

(1) When a superheterodyne receiver is tuned over a band of frequencies, a constant frequency difference must be maintained between the oscillator frequency and the incoming signal. When this difference is maintained properly, the oscillator tuning is said to *track* with the mixer tuning. The frequency of a tuned circuit will vary with changes in either the capacitance or the inductance. From the formula for resonance, it is clear that the frequency decreases as the square of the increase in capacitance or inductance. Since the oscillator generally is tuned to a frequency higher than that of the mixer, if variable capacitors are used in the oscillator and mixer circuits, the range of capacitance change of the oscillator must be restricted. The same change in capacitance in the oscillator as in the mixer would result in a larger oscillator frequency change at the higher frequencies, where a small change in capacitance is more significant. Restriction of the upper range of the oscillator tuning circuit is accomplished by a small capacitor in parallel with the main tuning capacitor. It effectively sets the minimum capacitance and consequently the maximum frequency of the circuit. Similarly, another much larger capacitor in series with the tuning circuit reduces the total effective capacitance so that the oscillator frequency maintains the required dif-



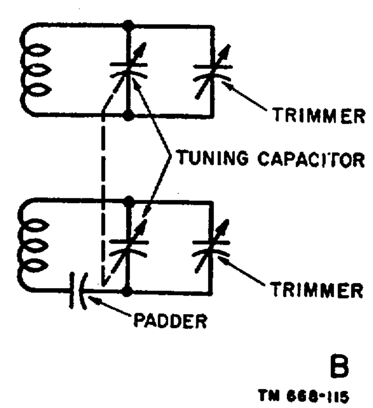


Figure 117. Tracking circuit and error curve.

ference from the mixer at the lowest frequency. By suitable adjustment of the inductance of the oscillator coil, the capacitance of the parallel capacitor (the *trimmer*), and the capacitance of the series capacitor (the *padder*), exact tracking can be obtained at three frequencies over the tuning range. Errors in tracking can be shown graphically, as in A of figure 117, where they are plotted against operating frequency. This type of error curve is typical for the circuit in B.

(2) When the inductance, rather than the capacitance, is varied (permeability tuned), the tracking is obtained either with variable inductances connected in series and parallel with the main inductance or with small trimmer and padder capacitors as before. Many double-conversion receivers require elaborate tracking arrangements. Special tuning capacitors may be used whose oscillator sections have differently shaped plates as compared with the mixer and r-f section. In this way perfect tracking can be obtained over the entire range without complex adjustment. This arrangement is not suitable, however, in a multirange receiver, since the frequency difference required between the oscillator and the mixer becomes proportionately smaller relative to the increased operating frequency. A coil can be wound with variable-pitch winding so that permeability tuning will give a constant frequency difference between oscillator and mixer. This method is used widely in the first i-f section of double-conversion receivers, with ganged-capacitor tuning in the second i-f section.

64. Mixer Tubes and Circuits

a. Diode Mixers.

(1) Diode mixers are used at the extreme upper end of the v-h-f range, where ordinary triode tubes are unsatisfactory. The diodes used have extremely small internal dimensions with close spacing between plate and cathode. A typical circuit using such a mixer is shown in figure 118. Diode mixers function not only with the oscillator fundamental, but also with harmonics. Therefore, oscillators operating at lower frequencies can be used. From the diagram, it is apparent that there is relatively little isolation between the oscillator and the r-f and antenna circuits, causing considerable oscillator radiation.

(2) The voltage from the oscillator is rectified, causing d-c to flow in the plate circuit. Optimum operation usually is obtained with d-c currents of approximately 200 to 500 microamperes. The parallel resonant circuit in the cathode is tuned to the oscillator fre-

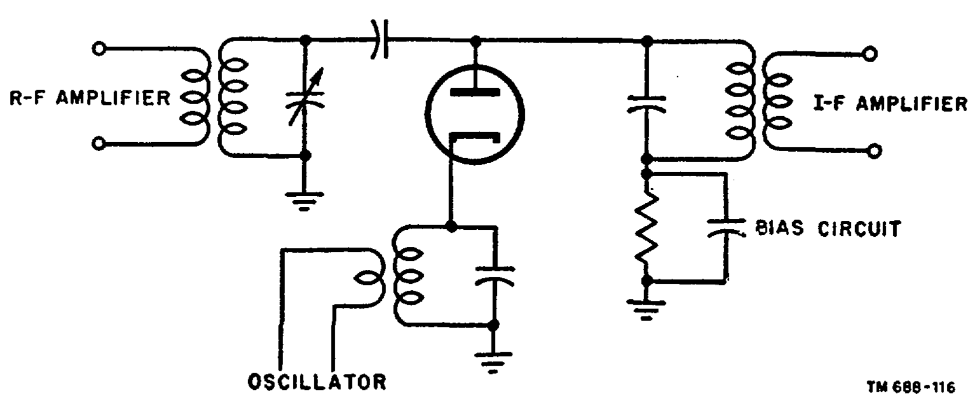


Figure 118. Diode mixer.

quency, with the capacitor made sufficiently large to act as a cathode bypass at the r-f signal frequency. The heater-cathode capacitance in most tubes is sufficient at high frequencies to render the oscillator injection circuit inoperative if not isolated. Therefore, r-f chokes always are required in the heater circuit with this type of mixer.

(3) Since the diode cannot have any conversion gain, the noise figure of this type of mixer depends entirely on the i-f amplifier which follows it. The diode produces a shot-effect noise, and noise, of course, is introduced by the oscillator injection. Therefore, the over-all noise figure of this type of mixer is not as good as in a crystal diode, where no filament and consequently no shot effect are involved.

b. Triode Mixers.

(1) Two triode mixer circuits are shown in figure 119. In A, the oscillator signal is injected along with the r-f signal at the grid; in B, cathode injection is used. As far as the performance of the mixer is concerned, there is little to choose between the two methods of oscillator injection, except that the grid loading with cathode injection is slightly greater. However, cathode injection gives better oscillator stability, since the load presented to the oscillator has a low impedance. The oscillator signal can be taken from a low-impedance point where varying load does not affect the operating frequency.

(2) The conversion transconductance is about 28 percent of the transconductance of the same tube operated as triode amplifier. If the transconductance at zero bias is reasonably high, the equivalent noise resistance of the tube is low, and the over-all noise is low. In general, the noise figure for a triode mixer is about the same as that for a single pentode amplifier. By the application of proper feedback, however, the noise from the oscilla-

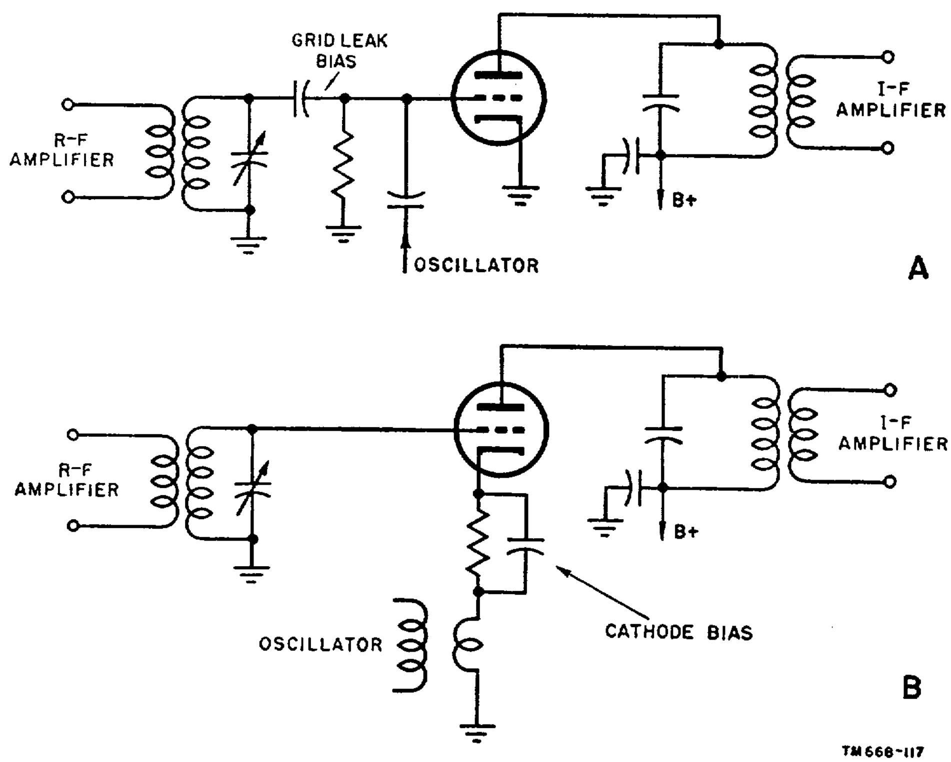


Figure 119. Single-ended triode mixers.

tor can be reduced so that the overall noise is no greater than that for the same tube used as an amplifier. It is possible to connect the tube as a grounded-grid mixer with cathode input and in this way obtain operation at higher frequencies than would otherwise be possible.

(3) If the operating frequency and the intermediate frequency are close together, the plate-load impedance becomes sufficiently great to permit tuned-grid tuned-plate oscillation. This causes instability, and the stage must be neutralized by conventional means. Furthermore, the capacitive plate load reflects a considerable resistive load across the mixer input, reducing the gain and the image rejection.

c. Dual-Triode Mixers.

(1) The push-push grounded-cathode circuit is used widely at v-h-f and functions effectively up to 600 mc. The oscillator signal usually is injected into the grid by means of a small inductive coupling loop (fig. 120). The local oscillator and input signal are completely cancelled in the plate circuit, which improves the signal-to-noise ratio and adds to the stability. The actual operation is identical with that of the balanced modulator, which it strongly resembles.

(2) Another dual-triode mixer that is useful at high frequencies is the cathode-coupled circuit shown in figure 121. The oscillator signal is injected through a cathode-follower section. The oscillator is isolated from the input circuit and from the mixer tube, which results in improved stability. There is almost complete elimination of any tendency of the oscillator frequency to shift with variation in the strength of received signal (pulling), as most other mixers would do with poor oscillator signal-grid isolation.

d. Pentode Mixer.

(1) The pentode shown in figure 122 is one of the most frequently used mixers in f-m equipment for the v-h-f

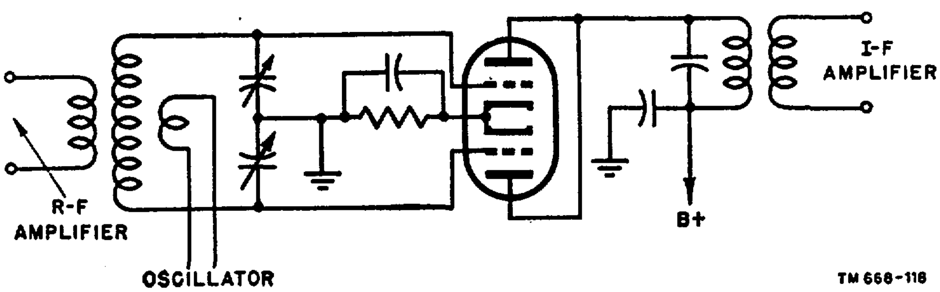


Figure 120. Push-push triode mixer.

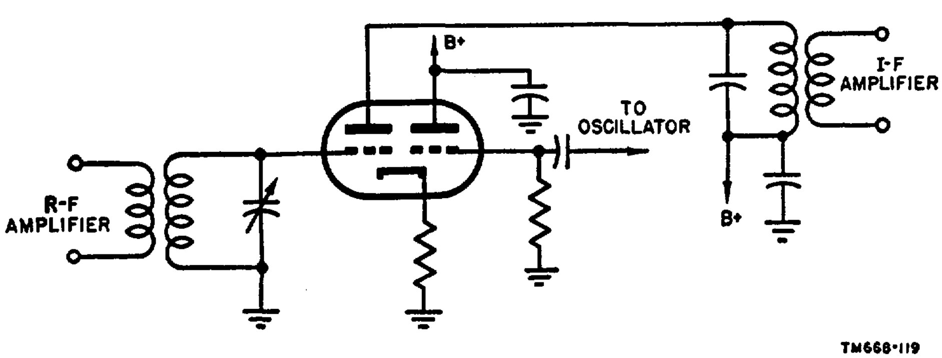


Figure 121. Cathode-coupled mixer.

band. At low frequencies, where the screen grid is effective, the mixer provides good isolation between the input and output circuits. This means reduced input loading and elimination of possible instability. The oscillator and signal voltages usually are applied to the signal grid. In this way, a noise figure is obtained which exceeds that of a normal pentode amplifier, but which is much lower than in any of the multigrid mixers.

cathode inductance, and cathode injection will lower the voltage gain of the input circuit and also the noise performance. The stability of the oscillator, however, is improved at very-high frequencies, where a low-impedance oscillator load is needed. Unless the oscillator and mixer are loosely coupled, interaction and pulling become severe. Interaction of the oscillator and the signal is greatest when they are both on the same grid.

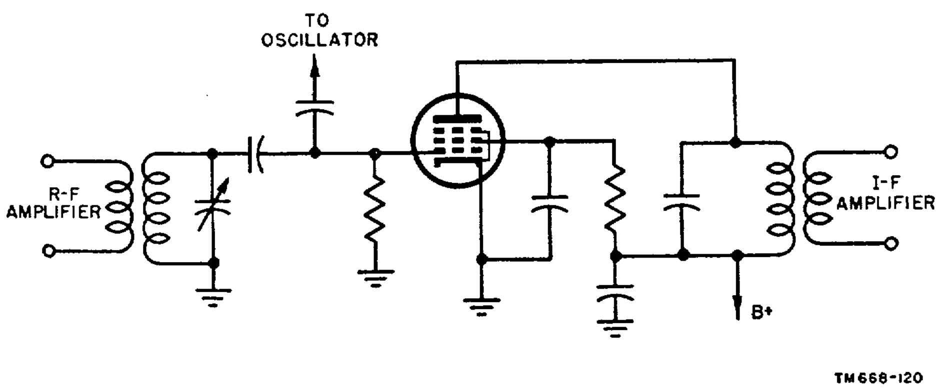


Figure 122. Pentode mixer.

(2) The pentode has an extremely high conversion transconductance and permits high voltage gain in the mixer stage. The equivalent noise voltage produced by the tube is twice that of a triode mixer of the same transconductance. Because of the high obtainable transconductance of pentodes, the over-all performance can exceed that of some triodes. Since the triode has a certain amount of coupling between grid and plate circuits, it is at a disadvantage compared with the pentode. At the signal frequency, the i-f circuit is capacitive, and this, because of Miller effect, results in a reflected low resistance in the grid circuit. The screen in a pentode effectively stops this loading.

(3) With a pentode, cathode injection of the oscillator signal is possible, but this mode of injection will increase the effective cathode inductance. Since the input load is proportional to the Similarly, oscillator radiation becomes a greater problem; however, the high transconductance permits the use of small oscillator voltages, and radiation is not as great a problem as in a triode.

e. Multigrid Mixers.

(1) In superheterodyne receivers, many different tubes have been developed solely for use as mixers, but only a few of these are usable at v-h-f. In general, all multigrid mixers, which have more grids than a pentode, are divided into two classes—those in which the r-f signal is placed on the inner (control) grid, and those where the local-oscillator signal is on the inner grid. In the first category are the pentagrid mixer and the heptode. In the second are the pentagrid converters, which include the hexode and the octode. In addition to these tubes, there are those with a triode local-oscillator section in the envelope with

a pentode mixer. These do not differ from the separate mixers and oscillators covered in the preceding paragraphs.

(2) A cross section of a typical pentagrid mixer is shown in A of figure 123. The oscillator voltage is applied to the oscillator grid, and the r-f signal is applied to the signal grid, which is the third grid out from the cathode. The signal grid is shielded from the oscillator grid by a screen grid. The screen grid is connected to a second screen between the signal grid and the plate which serves to isolate the signal grid from the plate. A conventional suppressor follows the second-screen grid. A typical circuit using this tube is shown in B. It has fairly high gain, because the internal construction permits a high conversion transconductance.

(3) Because of the large number of grids, the noise performance is poor, but this is largely offset by the high gain. With suitable feedback in the cathode circuit, the noise figure can be brought down to that of a pentode. The major fault arises from coupling between the signal grid and the oscillator grid through the space charge of the tube. This reduces the gain and the conversion transconductance. A small amount of capacitive coupling exists even though a screen is placed between the two circuits, and this adds to the interaction between the oscillator and mixer sections. The effect of the capacitive coupling is opposite to that of space-charge coupling, and if a small capacitor is placed between the oscillator and the signal grids to add to the normal capacitance, the space-charge coupling can be neutralized at high frequencies. This improves the performance somewhat. The tube loads the tuned circuit in the signal grid negatively so that the performance of that circuit actually is improved. However, during the nega-

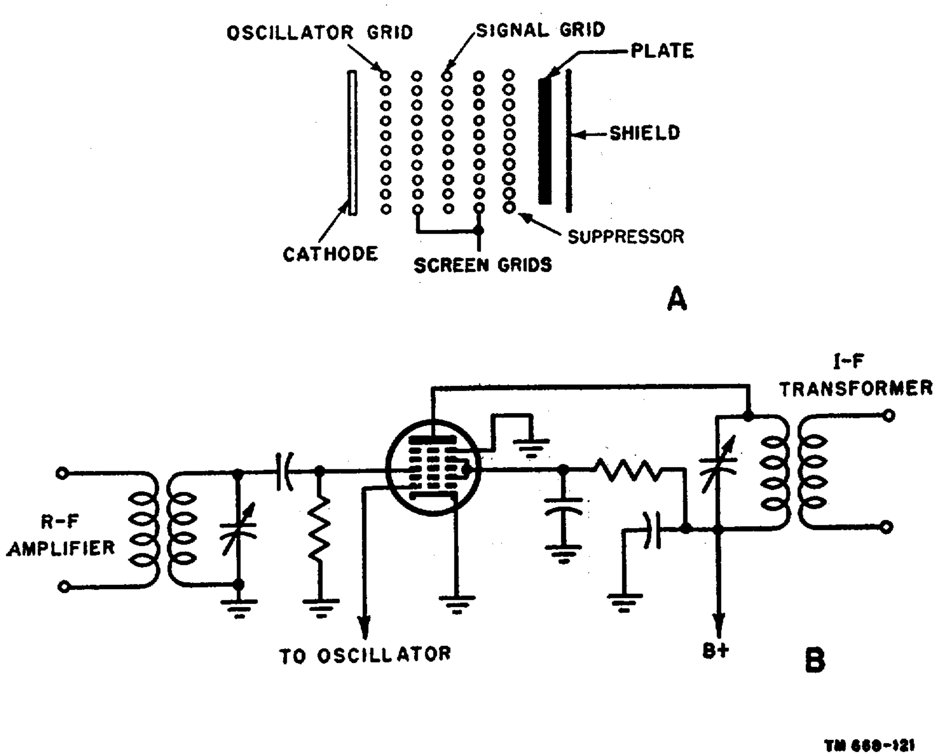


Figure 123. Pentagrid mixer.

tive portion of the oscillator cycle, the cathode can swing more negative than the signal grid. When this occurs, the signal grid draws current unless the oscillator grid is sufficiently negative to cut off the cathode current entirely. Therefore, suitable values of grid-leak resistance must be used to develop a high negative bias, or the oscillator voltage must be increased. The tube is used more often as a converter than as a mixer.

65. Converters

a. Converters are essentially mixer tubes where the oscillator grid in conjunction with the electron stream and the plate or screen acts as a self-contained oscillator. The oscillator, therefore, does not require a separate tube. The pentagrid converter is the principal type used, and a typical circuit is shown in A of figure 124. In a converter, the requirements of oscillator stability are most severe. Only those types can be used that have an internal construction which will allow the frequency of oscillation to be stable. The mixer part of the tube is conventional, but the oscillator consists of the inner grid, the cathode, and the screen grid. The screen acts as the plate, so that local oscillation takes place between elements forming a triode.

b. The converter is used widely in the second mixer-oscillator stage of double-conversion receivers, with a crystal oscillator as shown in B. It is a highly stable circuit with good gain at the low frequencies involved. The crystal oscillator is an ultraudion type, with the crystal acting as a tuned circuit between the screen and the control grid of the oscillator section, the screen acting as a plate. At the higher frequencies, the gain of the converter falls off unless sufficient capacitance is added between the oscillator and signal grids to neutralize the effect of space-charge coupling.

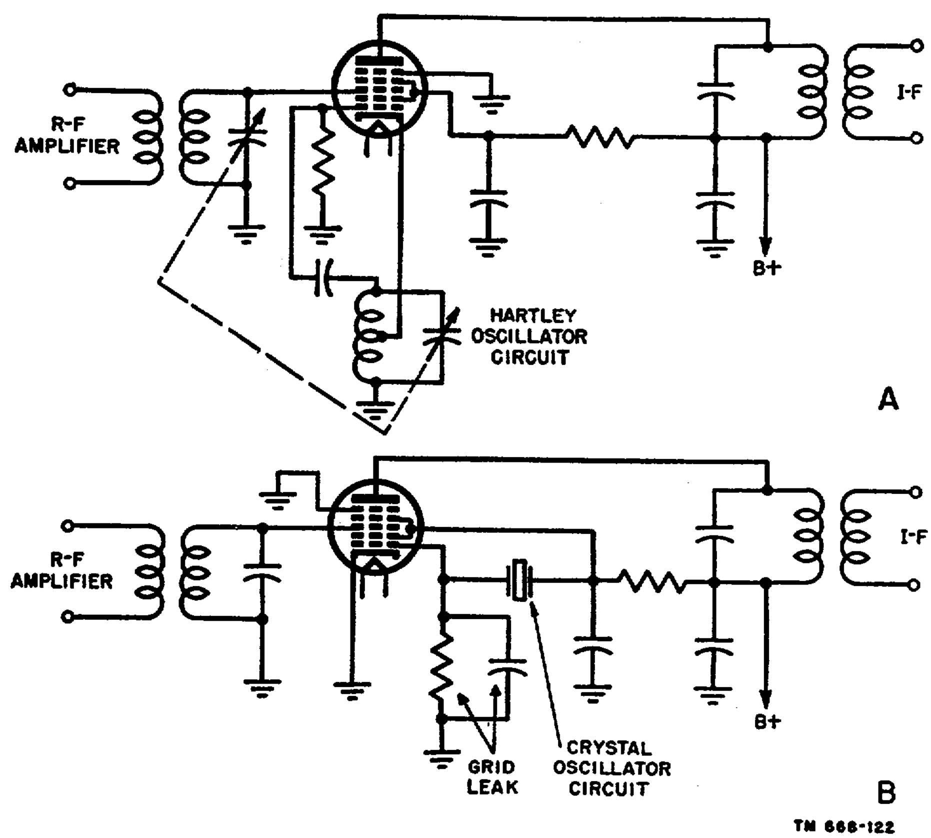


Figure 124. Pentagrid converters.

66. Crystal Mixers

a. General. At extremely high frequencies, vacuum tubes no longer function satisfactorily as mixers, and germanium or silicon diodes are used. The mixer circuit (fig. 125) is nearly the same as that of the diode mixer, although no bias or filament voltages are needed. The crystal operates because there is a much higher resistance to current passing in one direction than in the other. Therefore, the crystal is essentially a nonlinear device.

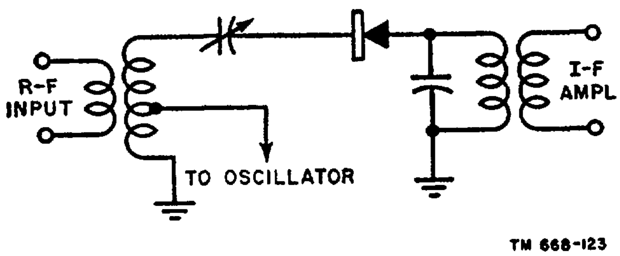


Figure 125. Crystal mixer.

b. Oscillator Injection. The performance of the crystal mixer depends on the uniformity and amplitude of the local-oscillator voltage injected. The oscillator voltage affects the matching of the r-f and i-f circuits and consequently the over-all noise performance. The average input impedance is approximately several hundred ohms, and the output impedance is about the same value, or slightly lower. However, they are not a constant and both input impedance and output impedance decrease as the rectified current produced in the crystal by the local oscillator increases. If ordinary methods of oscillator coupling with small capacitors are used, the injection of oscillator current varies widely, consequently changing the impedance. This affects adversely the noise performance and the conversion efficiency. Therefore, to provide uniform oscillator injection, equalizers are inserted in the circuit.

c. Conversion Efficiency. A crystal produces no voltage gain. However, the conversion loss is slight with well designed crystals. The effective loss in signal-to-noise ratio, when measured in terms of noise figure, also includes a factor resulting from the *excess temperature noise,* which is the additional oscillator noise measured by the rise in temperature needed to produce the same amount of noise with the same impedance level at the crystal input. The conversion efficiency is relatively high if sufficient oscillator injection is available. Power of approximately 500 microwatts usually is sufficient. This is much lower than that required by a vacuum-tube mixer.

d. Frequency Response and Noise Performance. The frequency response of a crystal is practically uniform from the low-audio frequencies up to the superhigh-frequency range. Germanium crystals are capable of withstanding higher voltages, whereas the silicon crystals have less noise and conversion loss. Since the crystals are used in ranges where r-f amplifiers are not practical, the conversion efficiency must be as high as possible with little noise. The over-all noise performance of an f-m receiver using a crystal mixer is determined almost entirely by the noise figure of the first i-f amplifier. In this respect, the i-f amplifier plays almost the same role as the r-f amplifier does at lower frequencies. It is necessary to use a high intermediate frequency with a crystal mixer to minimize oscillator radiation and excess temperature noise.

e. Crystal Operation at High Frequencies. The conversion transconductance of a crystal cannot be specified by a simple constant, since it depends on frequency and oscillator excitation. Because the spacing is extremely close, the capacitance between the fine wire in contact with the crystal surface in the holder is considerable despite the small area. No reliable prediction can be made of performance at high frequencies from measurements made at low frequencies. At these high frequencies, ordinary coils and capacitors are not satisfactory as circuit elements, so that it is common practice to use sections of transmission line, or waveguide. However, the operation of the crystal as a mixer is the same as any other nonlinear device, and at high frequencies, the over-all performance obtainable with good crystals approaches that obtainable at lower frequencies with vacuum tubes.

Section IV. OSCILLATORS

67. L-C Oscillators

a. General. The local oscillator of a superheterodyne f-m receiver generally operates at a high frequency. In a double-conversion heterodyne, however, the second local oscillator operates at a much lower frequency. The stability of the entire receiver is determined by the h-f oscillator since any change in oscillator frequency changes the effective tuning of the receiver. The principal requirement for high-frequency oscillators in f-m receivers is a high degree of stability in respect to thermal, mechanical, and electrical variations.

b. Lumped-Constant Circuits.

(1) Most oscillators used at very-high frequencies tend to be unstable unless great care is taken in the construction of the parts which determine the frequency. This instability stems from the large changes in reactance that take place with small changes in capacitance or inductance. The most satisfactory circuits are those in which the effect of the vacuum tube on the frequency of oscillation is least. This usually is accomplished through the use of the conventional Colpitts oscillator (fig. 126) or some modification of it. The conventional Colpitts oscillator in A obtains feedback through two capacitors in series, which are also effectively in parallel with the tube interelectrode capacitances. Since the interelectrode capacitances are small, large values of feedback capacitance make the effects of tube capacitance negligible.

(2) A maximum value of series capacitance, however, is determined by the necessary load impedance to be developed by the tank circuit in order to sustain oscillation. To avoid some of the disadvantages of using relatively low values of capacitance, the circuit in B often is used. The series capacitors are a part of the tank circuit, with a smaller series capacitor used to tune the oscillator. This arrangement permits stability over a wide range of variation in either tube or load conditions. Both circuits, however, have the disadvantage of a tuning range restricted by the size of the tuning capacitors. Therefore, even though stability is lower, conventional Hartley or Armstrong circuits are used where wide tuning range is necessary.

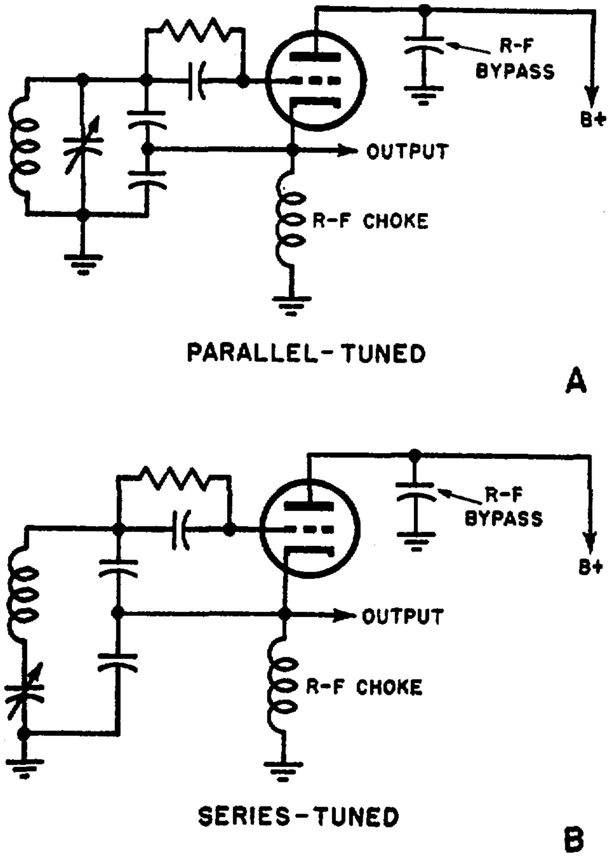


Figure 126. Two versions of Colpitts oscillator.

c. Causes of Drift.

(1) The causes of drift in lumped-constant oscillators operating at high frequencies can be classified electrically, mechanically, and thermally. Electrical drift takes place when the plate voltage, filament voltage, or any other tube potential changes. For most oscillators, the supply voltage enters into the determination of the frequency of

oscillation, although simplified theories of oscillator operation usually do not take this fact into account. This is because a vacuum-tube oscillator is a device whose characteristics vary with the supply voltage. The variations can be corrected by suitable voltage regulators for the plate and filament supplies. Mechanical drift arises when inductors and capacitors change their physical dimensions with vibration. This usually is overcome by using very rugged components and shock-mounting the unit.

(2) Thermal drift takes place because the values of capacitance or inductance change as the dimension of the components change with temperature. Low temperature-coefficient ceramic materials, stretched wire coils, temperature-compensating capacitors, and similar devices reduce the drift. Associated with thermal changes are humidity variations. Moisture-laden air has a different dielectric constant from that of dry air, and the capacitance of air capacitors changes with variations in humidity. To overcome this, a heating device can be located near the oscillator tank to keep the air dry, or the tank circuit can be hermetically sealed. Permeability tuning, which is much less susceptible to humidity, also can be used where other methods are not satisfactory. In all instances, the instability can be reduced by lowering the frequency of the oscillator and then using frequency multipliers to raise it to the desired value.

d. Distributed-Constant Oscillators. At frequencies that are too high for ordinary lumped-constant oscillators, the tuned circuits are made of sections of transmission line. A shorted coaxial line serves as an inductance if its electrical length is less than a quarter-wavelength at the operating frequency. These transmission-line sections have very high Q and, if properly designed, are stable mechanically. Receiver oscillators using transmission lines tuned with variable capacitors are common in v-h-f equipment. The inductance is distributed along the entire length of the line, and therefore is called a distributed constant. The operation of these oscillators is the same as for those used at low frequencies except that the tuned circuit does not have a lumped inductance.

e. Second-Conversion Oscillators. In the second conversion circuit of double-conversion superheterodynes, the local oscillator can be of any conventional type. In some double-conversion receivers, the tuning is done in the second oscillator because it is easier to obtain stability at low frequencies.

68. Crystal-Controlled Oscillators

a. General. Crystal oscillators are used in f-m receivers because of their extremely high stability. Crystal oscillators used in second-conversion circuits operate at low frequencies with output at the frequency of the crystal. When crystals are used in the first-conversion circuit, the crystal oscillates at its fundamental frequency, but harmonic output is obtained through a harmonic generator. There are also circuits that cause the crystal to oscillate directly on a mechanical harmonic. For low-frequency circuits, the choice of circuit depends on its simplicity and economy of parts. In the high-frequency converter, the requirements are more complex since there is conflict between requirements for high harmonic output, low subharmonic output, stability, and low-crystal current.

b. Low-Frequency Crystal Oscillator.

(1) The crystal is the equivalent of a high Q resonant circuit and controls the frequency of an oscillator precisely. Each crystal oscillator circuit has its counterpart in an L-C oscillator. For example, the triode-crystal oscillator in A of figure 127, is the equivalent of the tuned-plate, tuned-grid oscillator. In B, the crystal oscillator circuit is the equivalent of the ultraudion oscillator. In both circuits, the output voltage is at the frequency of the crystal. The tuned-plate, tuned-grid arrangement is used rarely because precise tuning of the output circuit is necessary to maintain oscillations. The ultraudion circuit oscillates because of the feedback provided by the voltage

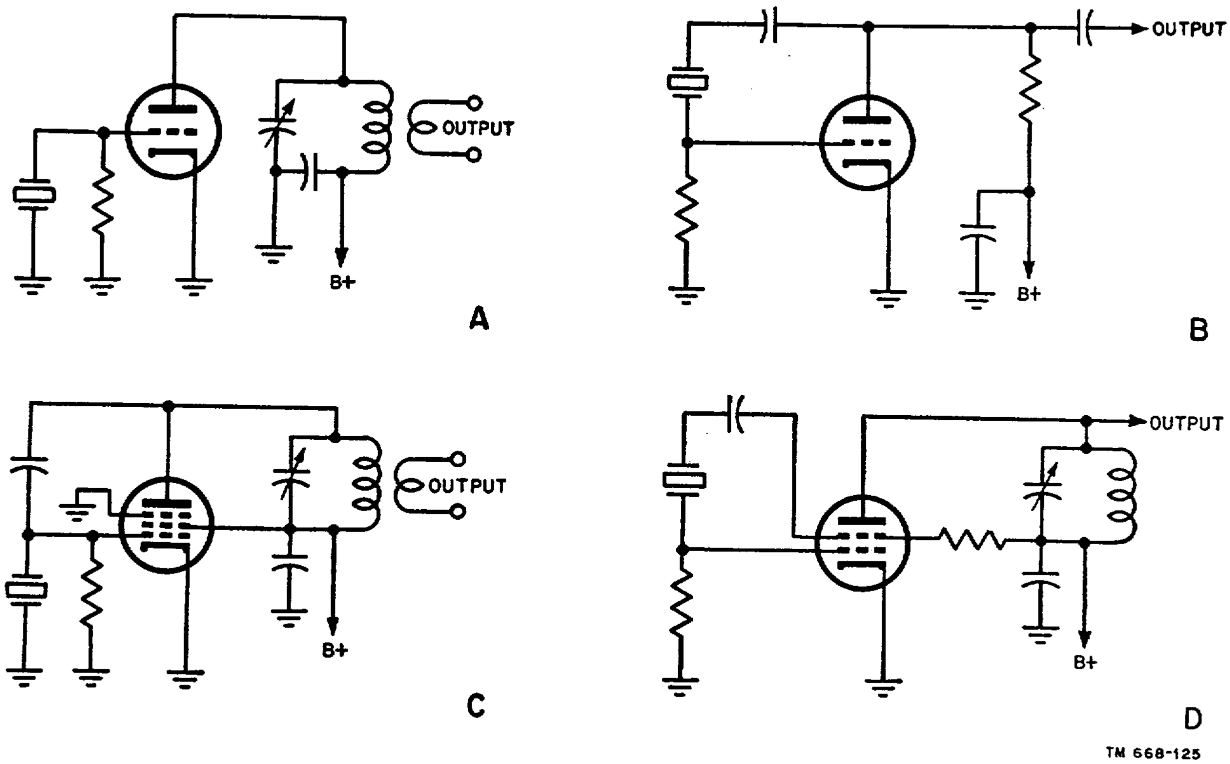


Figure 127. Low-frequency crystal oscillators.

divider formed by the capacitance from the grid to the plate and the cathode. It is perhaps the simplest of all crystal oscillators, since it requires only resistors and capacitors.

(2) Other crystal oscillators whose output voltage is of the same frequency as that of the crystal are shown in C and D. These are designed to operate with tetrodes and pentodes. The circuit in C is the same as the tuned-grid, tuned-plate in A, except that the triode is replaced with the pentode. The circuit in D is similar to the ultraudion in B, but it uses a tetrode.

c. Harmonic-Oscillator Circuits.

(1) Where the local oscillator of the f-m superheterodyne operates at a very-high frequency, it is not possible to use a crystal that oscillates directly at this frequency. One solution for obtaining output in the v-h-f range is to use an oscillator which generates a strong output on harmonics of the crystal fundamental, and then to use separate frequency multiplication to obtain the desired frequency.

(2) Figure 128 illustrates three harmonic-oscillator circuits for pentodes. The cathode, control grid, and screen grid form a triode oscillating at the crystal frequency. These oscillations are rich in harmonics and the plate circuit is tuned to the desired harmonic. A high impedance is presented to the selected harmonic and a considerable output voltage is developed across this load. The three circuits differ in the basic type of oscillator; that is, in control grid-screen, grid-cathode connections. In A, the crystal is inserted between the control grid and the screen grid. In B, it is inserted between the cathode and control grid.

(3) In C, a resonant circuit in series with the cathode is tuned midway between the crystal frequency and the second harmonic. Therefore, it is inductive at the crystal frequency. In conjunction with the capacitance between grid and

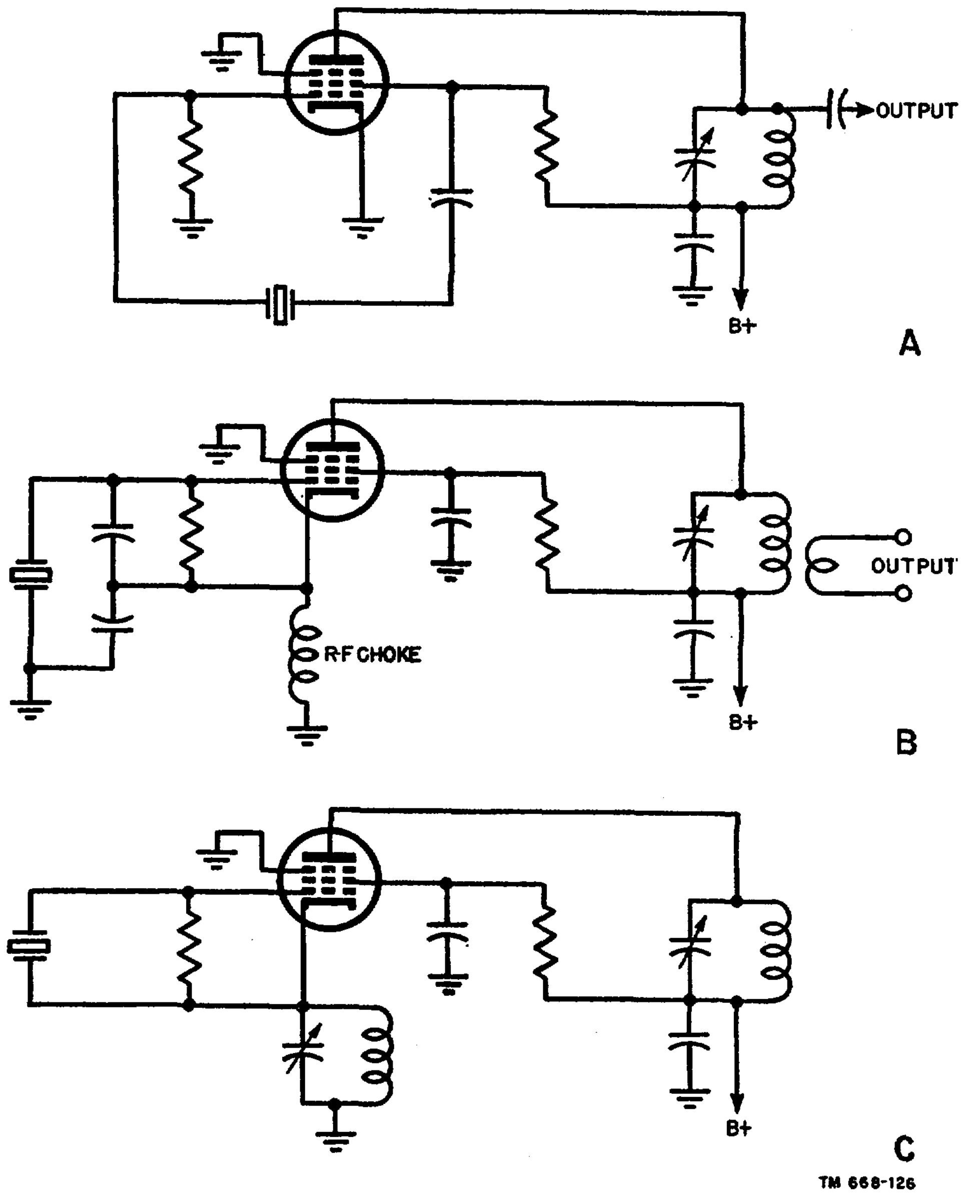


Figure 128. Harmonic crystal oscillators.

cathode, it acts as a voltage divider providing the required feedback for oscillation. Of these three circuits, that of C has the greatest harmonic output and the highest stability when it is used with a well-screened pentode. The second circuit is practically independent of variations in tube characteristics, if the feedback capacitors are large enough.

d. Overtone Oscillators.

(1) The output of all of the harmonic oscillators contains components at frequencies other than the fundamental. Unless they are separated from the mixer by a sufficient number of tuned circuits, serious difficulty with spurious responses can arise. These disadvantages can be overcome with special circuits and crystals. Instead of oscillating at their fundamental frequency, crystals can be made to oscillate on other frequencies very close to odd harmonics of their fundamental.

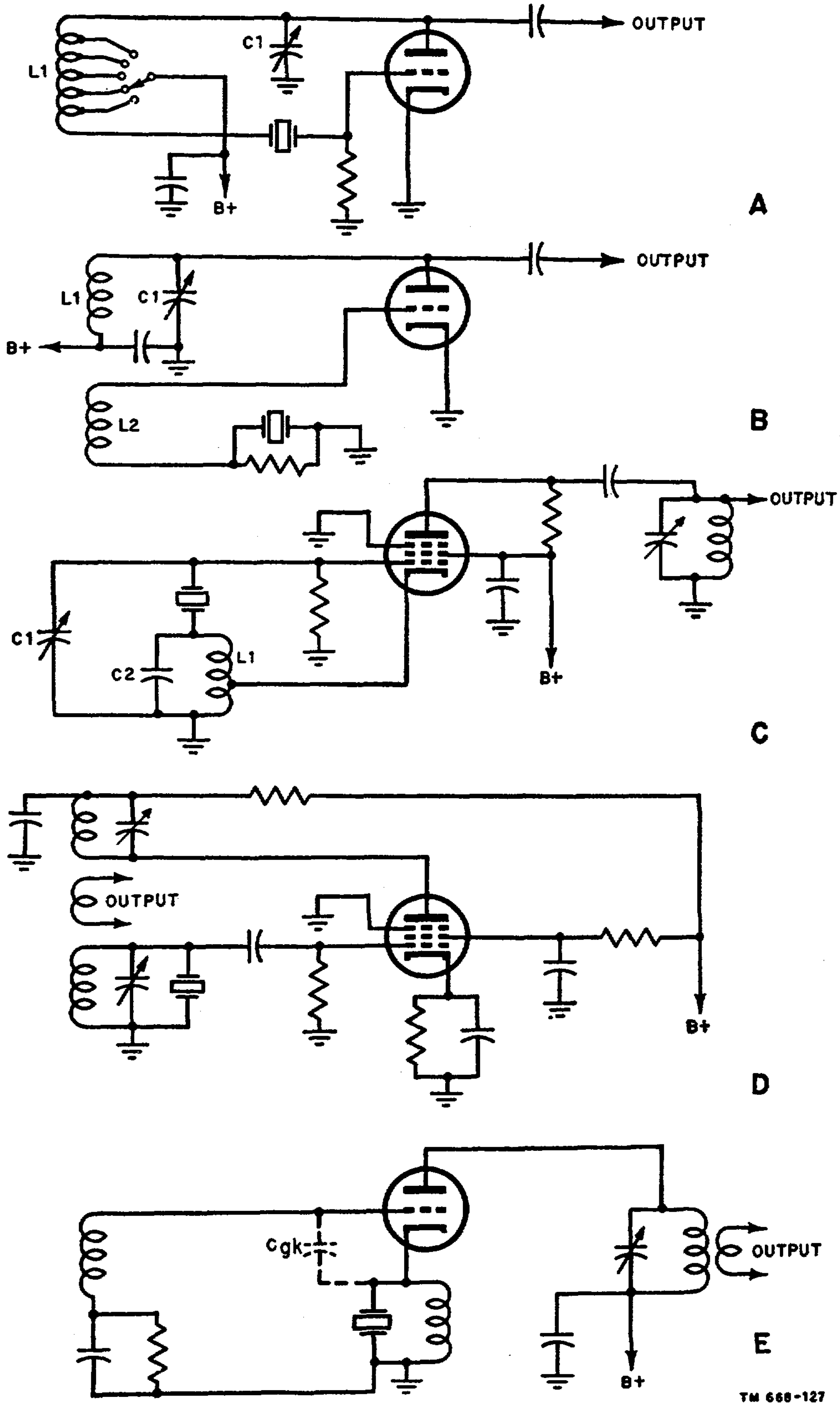


Figure 129. Overtone oscillators.

These frequencies are its mechanical harmonics. Oscillators using this principle are called *overtone oscillators*. Special types of circuits must be used for this purpose. Their advantage lies in negligible output at frequencies lower than the desired frequency, which reduces the number of possible spurious responses considerably.

(2) All of the overtone oscillator circuits (fig. 129) incorporate some form of frequency-selective feedback unlike that in a normal crystal oscillator. An additional resonant circuit is used to feed back energy at the frequency of the desired overtone, so that oscillation takes place only at the desired frequency. The frequency-selective feedback must be adjusted carefully, so that oscillation is caused only by the crystal and not by the feedback circuit.

(3) In A, a simple ultraudion triode oscillator is modified so that the frequency of feedback is controlled by the tuned circuit formed by *L*1 and *C*1. The amount of feedback is varied by changing the tap on *L*1. A second version of the same oscillator is shown in B, where the amplitude of the feedback is controlled by the coupling between *L*1 and *L*2. These circuits are especially suitable for operation of the crystal on the third overtone, where the frequency of feedback is not very critical. Power output at the fifth overtone is poor, and the feedback adjustment becomes fairly critical when ordinary crystals are used.

(4) The overtone oscillator in C is capable of extremely stable operation, although it works only with crystals designed for overtone service. Feedback is controlled by the position of the cathode tap on *L*1 and by the setting of capacitor *C*1. The tuned grid circuit is set for resonance at the frequency of the desired overtone. Therefore, the capacitance of the crystal and its holder is also part of the resonant circuit, which is similar to that of the Hartley oscillator. This circuit has good output, especially at higher overtones, and the plate circuit can be tuned to a harmonic of the overtone, producing further frequency multiplication.

(5) An overtone oscillator which uses a high-gain pentode is shown in D. Feedback is obtained by magnetic coupling. The crystal actually is resonated slightly above the desired overtone. This makes the entire grid circuit equivalent to a high-impedance parallel-resonant tank, which easily picks up a regenerative signal from the plate circuit and produces oscillation. At very high overtones, the capacitance of the grid circuit is sufficient to resonate with the inductance in the plate circuit. This circuit is capable of producing moderate amounts of output on extremely high overtones. Operation on the twenty-ninth overtone has been obtained. This represents an important saving of frequency multiplier stages, as well as freedom from spurious responses attributable to the subharmonics of doublers and triplers.

(6) The circuit in E is somewhat different from those above. The input and output circuits of the triode are tuned to the same frequency, although the grid is much more broadly resonant than the plate. This is because of the high ratio of inductance to grid-cathode capacitance. The crystal-holder capacitance is resonated in the cathode circuit with an inductor at the desired overtone frequency. The only frequency at which oscillation can take place is that at which the crystal goes through series-resonance, producing a low impedance from cathode to ground. To prevent self oscillation between grid and plate circuits, the inductance in the cathode circuit must be adjusted carefully.

Section V. TUNING AND FRONT-END CONTROLS

69. Front-End Design

a. Since few f-m receivers are designed to operate on a single frequency, means must be provided for tuning them over the required range. The mechanisms used, as well as the circuits are designed to operate in the r-f amplifier, mixer, and oscillator stages and are known as the *front end* of the receiver. In many instances, operation is required over a considerable portion of the v-h-f spectrum, and the requirements placed on the tuning circuits become severe. Some form of automatic tuning also is desirable.

b. Since all tuned circuits use inductance and capacitance in one form or another, all front ends are designed to vary one of these basic circuit elements. At frequencies low enough for lumped constants, variable capacitors or variable inductors are used. Different means are available for adjusting these elements, such as powdered-iron slugs, tapped coils, or variable capacitors. At higher frequencies, distributed-constant devices, such as transmission lines, are used. The length of the transmission line is made either electrically or mechanically adjustable. Where crystal control is used, switching circuits have been evolved to change from channel to channel.

70. Automatic Tuning Mechanisms

a. Detent Selector. One of the simplest of all automatic tuning selectors is the detent switch. This device enables a shaft used for tuning to be set repeatedly at the same position (fig. 130). The actual tuning usually is accomplished by a variable capacitor connected to the shaft. A gear and dial-drive arrangement permits selection of frequency groups comprising different channels, and also selection of individual channels within each group.

b. Mechanical Push-Button Systems. A device that rotates the shaft of a variable capacitor to preset positions is shown in figure 131. The buttons are attached to notched rods resting on gears. When the button is depressed fully, the rod pushes the gear to a point where it can rotate no farther. The gear, in turn, rotates a variable capacitor to the desired position.

c. Electrical Push-Button Systems. Electrical push-button systems are manually operated electrical switches which connect various trimmer capacitors in turn across the tuned circuit. The setting of these trimmers determines the frequency to which the receiver is tuned. Included in this arrangement is a mechanical

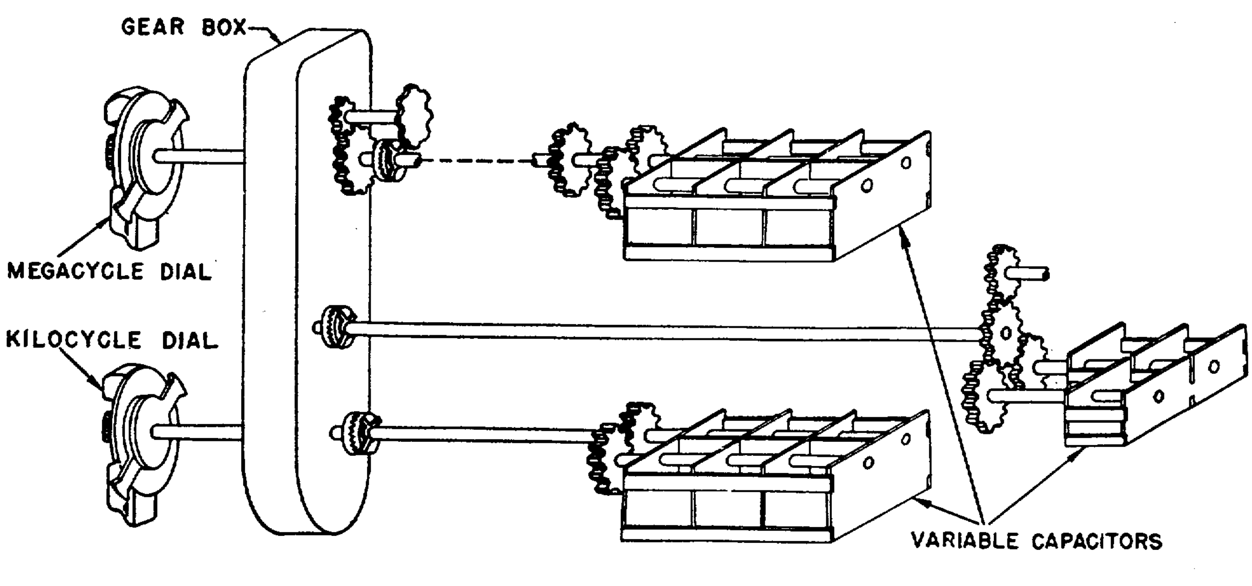


Figure 130. Detent mechanism.

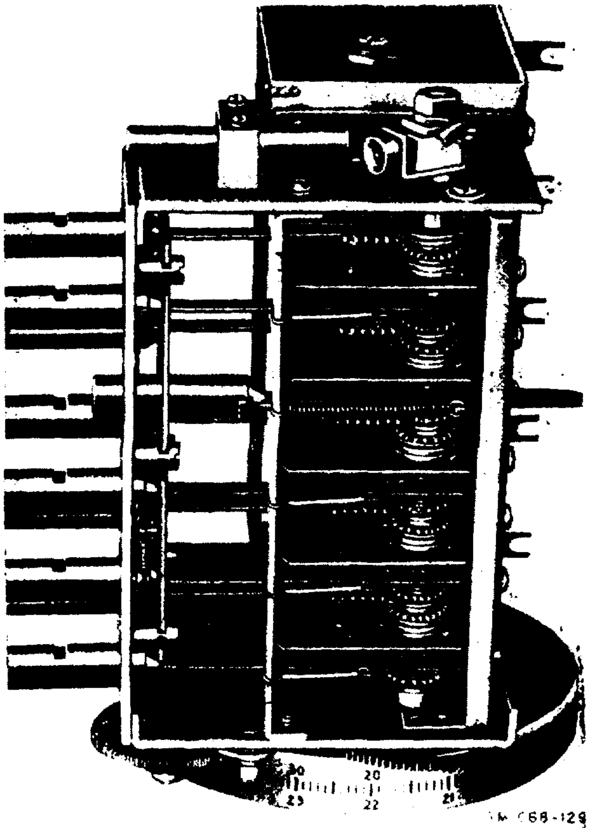


Figure 131. Mechanical push-button system.

latching gear that makes it impossible to depress two buttons simultaneously. In addition, the latching bar also disengages any button when another is pressed. The last button pressed stays engaged on the latch bar, indicating the channel in use. Because of excessive r-f losses in the complicated structure of the latching switch, it is not as satisfactory at v-h-f as are the mechanical positioning systems. However, up to about 100 mc, the performance is satisfactory if there is sufficient shielding.

d. Switch and Turret Tuners.

(1) A rotary selector switch used with preset coils to form a front-end circuit has many advantages over the ones previously mentioned. A rotary switch with several decks can connect in turn the appropriate reactors for up to approximately 12 channels. Each preset channel can be adjusted for optimum performance without much interaction from the other reactors if sufficient shielding is used. This system has good noise performance as well as low circuit capacitance, which is especially valuable at high frequencies. However, the limitation in the number of channels, as well as the necessity for very elaborate equipment, limits the use of this arrangement. A more flexible variant combines the switch tuner with a variable capacitor. The switch selects the appropriate coils and trimmer capacitors, and the variable capacitor provides fine tuning over the range of the particular coil.

(2) One of the most efficient of all the multichannel selectors is the so-called *turret* tuner. In this device, all of the channels have separate sets of inductors and capacitors which are mounted in a drum, or turret. The efficiency is high because the leads between the tuning elements and the tubes can be made short, permitting high circuit inductance and low over-all capacitance, which in turn means high tuned-circuit gain. As with all preset channel tuners using fixed electrical elements, it is difficult with the turret tuner to change a whole group of channels. This disadvantage can be overcome by providing an auxiliary variable capacitor or inductor which permits tuning through a considerable range.

71. Manual Tuners

a. Fixed-Frequency Equipment. In receivers where there is no need to change the band of frequencies, and where the total tuning range is small, separate coils and capacitors which are tuned manually are provided for each stage. An ordinary socket serves for the crystal oscillator, and different crystals are plugged in. The tuning of the r-f amplifier and mixer stages is accomplished with screw-driver-adjusted trimmer capacitors. At higher frequencies, where coils and capacitors no longer provide the proper amount of reactance, small setscrews which vary the length of a tuned transmission line are used in a similar fashion. Other arrangements, less frequently used, involve devices that are

combinations of inductors and capacitors and are varied with a setscrew.

b. Variable High-Frequency Equipment. The receiver shown in figure 132 has coaxial lines made of tubing. These are inclosed in square cans and are tuned electrically by a ganged variable capacitor. Another arrangement has been tried, using parallel-tuned lines that are actually parts of the capacitors. So-called *guillotine* tuners, which are essentially variable inductances, also are used. The parallel line is shorted by a bar, giving the appearance of a guillotine. Another tuned-line arrangement uses flat-plate transmission lines, with variable capacitors formed by small metal disks for adjustment of resonance range.

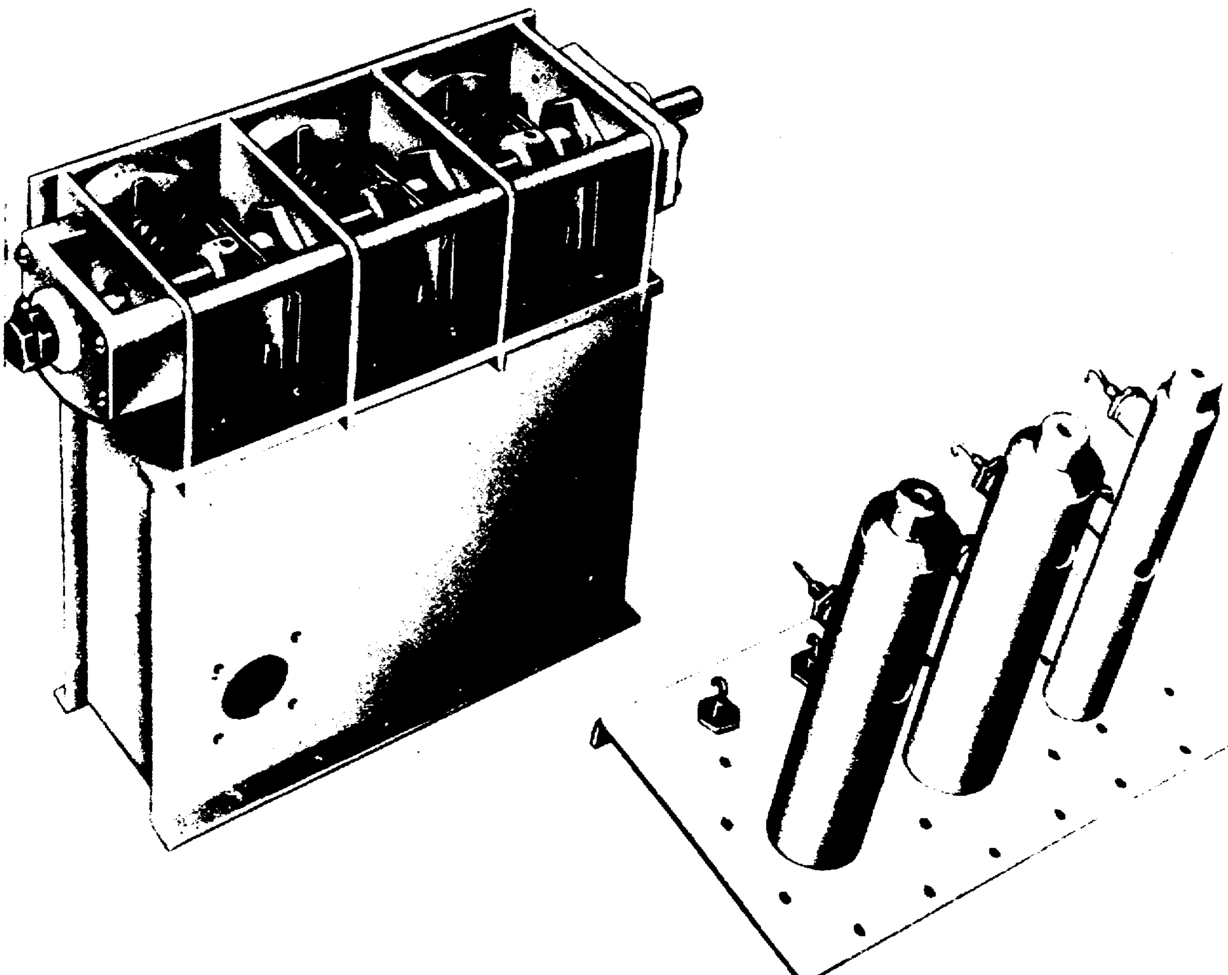


Figure 132. Manual tuning of coaxial lines.

Section VI. I-F AMPLIFIERS AND LIMITERS

72. Function

a. General. The front end of the f-m receiver produces an output signal usually in the microvolt range, whereas the detector requires signals in the order of several volts for its operation. Therefore, the i-f amplifier must perform most of the voltage amplification. The i-f stages must amplify the signal between 100,000 and 1,000,000 times. They also must introduce sufficient selectivity to discriminate against stations operating in the adjacent channel, and yet

have response sufficiently broad that the outer side bands of the f-m signal are not distorted.

b. Selectivity. The selectivity of an i-f amplifier is a measure of the total response of all of its tuned circuits. To obtain the required gain with adequate bandwidth, two or three i-f stages usually are needed. The circuit of a three-stage i-f amplifier is shown in figure 133. Each of the eight tuned circuits from the output of the mixer to the input of the detector has a response curve of the familiar bell shape. The response of the entire amplifier is the product of the responses at each point of the curve for each tuned circuit. The broadness of the peak of the curve for each tuned circuit depends on the Q. If the Q is 100 at a frequency of 10 mc, the peak of the curve is 100 kc wide between points where the voltage falls off to .707 of the peak value, or 3 db down. In general, the response of a tuned circuit is given approximately by the relation,

$$\text{bandwidth in kc} = \frac{\text{center frequency in kc}}{Q}.$$

When two tuned circuits are used, the response of the circuit is sharpened. or example, the over-all bandwidth of the circuit is 100 kc, and the response, by definition, is 3 db down 50 kc to either side of the center frequency. Applying the output of this circuit to a similar one means that a signal 3 db down at the input is reduced another 3 db, or a total of 6 db. Neglecting the mutual inductance, the response at 50 kc from the center frequency for an amplifier with eight tuned circuits is

$$8 \times 3 \text{ db} = 24 \text{ db down.}$$

This illustrates how the bandwidth is narrowed by increasing the number of tuned circuits. As the bandwidth is narrowed, the selectivity is increased.

c. Selectivity Requirements.

(1) In low-frequency f-m equipment, the channel spacing between adjacent transmitters can be as low as 50 kc. To prevent interference between channels at the receiver, considerable selectivity is required; however, increasing the selectivity of a stage increases the amplification of that stage. Because of difficulties resulting from feedback and instability, there is a practical limit to the amplification that can be obtained from an i-f amplifier. It seldom is possible to obtain high selectivity in a single i-f section. For good image rejection a high i-f is needed, which further restricts the selectivity that can be obtained. The double-conversion receiver separates the i-f section into two sections operating on different frequencies and helps reduce instability. The high i-f permits the necessary image rejection, and the low i-f provides the necessary selectivity.

(2) For higher frequency operation, where wide-band f-m is used, double conversion is not needed, since the adjacent channel-rejection requirements are less severe. However, wide deviation means that the selectivity of the i-f amplifier must be sufficiently broad to handle the full swing and still provide high gain and adequate rejection of off-channel signals.

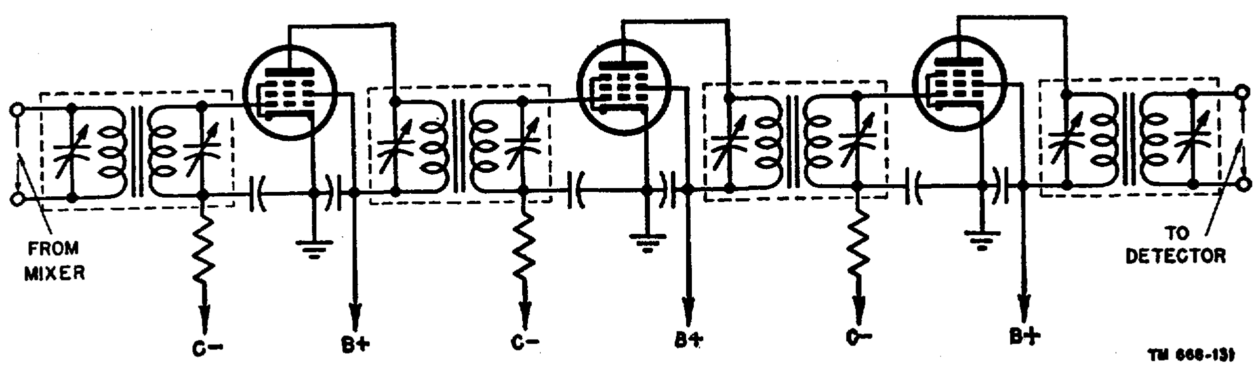


Figure 133. Three-stage i-f amplifier.

73. Tuned Voltage Amplifiers

a. Transformer-Coupled Amplifier. A transformer-coupled amplifier, such as a typical stage in the circuit of figure 133, has a selectivity characteristic that is dependent on the amount of coupling between the primary and secondary of each transformer. The response curves for various degrees of coupling in such circuits are given in figure 134. Since the f-m side bands extend for a considerable distance on either side of the center frequency, a broad flat response is desirable. Beyond this, a sharp decrease in response is necessary to attenuate adjacent channels. An ideal curve for the f-m circuit would be the curve in B, but the sharp corners and vertical sides cannot be attained with conventional tuned circuits. By selecting the proper amount of coupling, such as that of the curve where K, the coefficient of coupling, is equal to .015, an approximation to the ideal curve where K is equal to 1 can be reached in a large number of stages. The value of K is given by the following formula:

$$\text{Coefficient of coupling } K = \frac{1}{\sqrt{Q_p Q_s}}$$

where Q_p and Q_s are the values of Q for the primary and secondary, respectively. For example, suppose that the primary and secondary Q's are equal to each other and have a value of 66.7. Then, from the formula,

$$K = \frac{1}{\sqrt{66.7 \times 66.7}} = \frac{1}{66.7} = .015.$$

Such a circuit has the response characteristic indicated by that value of K shown in A of figure 134.

b. F-M I-F Transformers. The operation of the interstage coupling transformer determines the gain and selectivity of the stage. In general, these transformers must be adjustable so that the receiver can be tuned for maximum performance. The transformer can be tuned by variable capacitors across the fixed inductance of the primary and secondary, or the capacitors can be fixed and the inductance of the coils varied by means of powered iron slugs. The coupling of the coils is adjusted at the factory by setting the distance between the primary and the secondary. Low-frequency i-f transformers used in the second conversion section of double superheterodynes are similar.

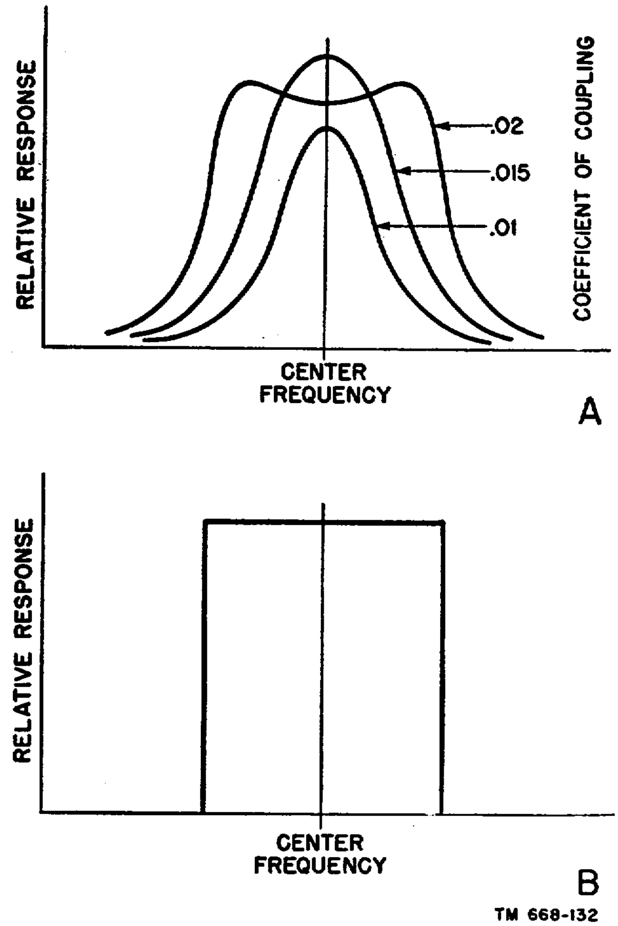


Figure 134. Response of i-f transformers.

c. Distortion in F-M I-F Stages.

(1) To gain the full advantages of f-m reception, the i-f system must have a selective response that does not introduce objectionable distortion into the side-band system. At the same time, it must have sufficient adjacent-channel rejection. If the selectivity curve is not symmetrical about the center frequency, considerable distortion appears in the output of the receiver because some side bands are amplified more than others. The parallel-tuned transformer, in addition to amplitude-response variation, introduces a certain amount of phase shift into the signal. If the phase shift is not uniform with frequency, the effect is somewhat like additional spurious

modulation of the signal. An ideal selectivity characteristic has a uniform phase shift throughout the band of frequencies that it passes.

(2) To correct distortion in practical circuits, it is necessary to make the response curve as flat as possible. Although the double-humped (overcoupled) curve in A of figure 134 provides sufficient bandwidth and steep sides, it has bad phase distortion. However, if a single-tuned circuit is added to this double-humped curve, the shallow depression in the center can be raised, giving a broad flat-topped characteristic with sharply falling sides. Alternate stages are used with either single-tuned circuits or undercoupled transformers and these fill in the depression caused by the overcoupled circuits of other stages. In the typical amplifier, eight sets of coils are contained in four transformers. If two are overcoupled and two critically coupled by the right amount, a curve can be obtained which closely approaches the ideal. Unfortunately, overcoupled circuits are difficult to aline. The tuning of the primary and secondary interact strongly with one another, with the result that the optimum response curve is difficult to obtain without special equipment and tuning procedure.

(3) Another way to achieve the same result is to couple all of the stages critically so that each response curve is essentially that of a single-tuned amplifier. Each stage then is detuned slightly above or below the center frequency (stagger-tuned). The successive stages each provide a portion of the desired flat top of the selectivity curve, and the sides of the curve are obtained by the two stages tuned farthest from the center frequency on either side. No two stages are tuned to exactly the same frequency; therefore, any tendency to instability and self-oscillation is reduced considerably. This method introduces phase distortion into the f-m signal.

d. I-F's for High Adjacent-Channel Rejection.

(1) Where extremely good selectivity is required, the standard double-tuned i-f transformer is not satisfactory. Even in double-conversion receivers there can be enough residual response in the adjacent channel to make reception difficult when a powerful local transmitter is operating while the receiver is tuned to a weak, distant station. Two methods designed to overcome this difficulty are the triple-tuned i-f transformers and the band-pass filter. The triple-tuned transformers have two ordinary windings inductively coupled to each other, but the output of the secondary is fed into a third parallel-resonant circuit through a capacitor. In this way, the coupling of the three circuits can be arranged for good gain and very sharp selectivity. A sufficient number of these stages provides good attenuation of the adjacent channel.

(2) The other method is more complicated, but produces better results. The output of the second mixer is fed into a band-pass filter that contains several tuned circuits. This filter produces a nearly ideal selectivity characteristic. Following the filter is a three-stage resistance-coupled i-f amplifier, which in turn drives the detector. The gain ahead of the filter is deliberately kept low, and the strong adjacent-channel station therefore cannot cross-modulate the weak station to which the receiver is tuned. The filter selects the weak station and rejects the strong one before there is any appreciable voltage gain. All of the voltage gain takes place in the resistance-coupled stages, which receive only the weak signal from the filter, the strong adjacent channel station being almost completely rejected.

74. Stability in F-M I-F Amplifiers

a. Transformer-Coupled Pentode Stage.

(1) If an f-m i-f stage is unstable, the

selectivity characteristic will not be symmetrical, and distortion will be high. Instability may be caused by feedback of energy from the output to the input, either in a single stage or over the chain of amplifiers. This instability usually does not exist on both sides of the pass band of i-f transformers at one time, since the feedback can operate over only a narrow range of frequencies. Therefore, assuming that all tendency to oscillate takes place at or near the center frequency, the amplification will be greater on the side of the response curve that tends toward oscillation. Oscillation may occur in voltage amplifiers at points far removed from the operating frequency, but these are not of immediate concern. An unstable or regenerative amplifier not only has bad selectivity characteristics, but will also respond more strongly to changes in supply voltages, tube characteristics, or signal input. For the best operation of an i-f amplifier, all voltage feedback must be reduced to a minimum, so that interchanging tubes and circuit components for repair does not produce undesirable results. Although it is possible to reduce the tendency to oscillate by reducing the gain of the stage, the loss of gain impairs the sensitivity of the receiver.

(2) The major cause of instability in pentode amplifier stages is the slight residual capacitance that exists between the grid and the plate of the tube, tending to cause tuned-plate, tuned-grid oscillation. The value of this residual capacitance is seldom above .005 $\mu\mu$f, and it exists even if the screen grid is perfectly grounded. This small capacitance is sufficient to promote oscillation when high-gain tubes are used with large values of inductance in the interstage transformers. The feedback can be neutralized as in a conventional triode amplifier, but the small capacitance makes the adjustment of any usual neutralization circuit difficult. The circuit in A of figure 135, shows one method of neutralizing this small capacitance by using a common bypass capacitor for the plate and screen circuit. If properly chosen, this capacitor forms a bridge circuit, as in B, consisting of the grid-plate capacitance, C_{gp}, the grid-screen capacitance, C_{gs}, the plate-suppressor capacitance, C_p, and the neutralizing bypass, C_n. At terminal 1, which is in the grid circuit, no voltage can be received from terminals 2 and 3, the output circuit, because it is canceled out in the bridge. At 10 mc, the neutralizing capacitance is about .002 μf for the typical pentode. The bridge does not

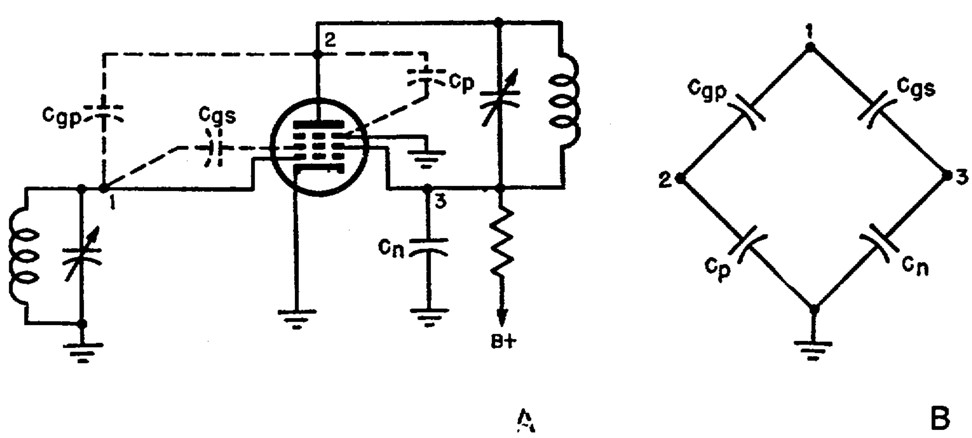


Figure 135. Neutralization of pentode i-f amplifiers.

provide perfect balance, but it materially reduces the tendency toward feedback through the grid-plate capacitance.

(3) Other types of feedback also limit the gain that can be obtained from the stage without oscillation. For example, since the impedance of the circuit is proportional to the inductance and also to the *Q*, increasing the inductance raises the impedance and the over-all gain. If the inductance is made large enough with the *Q* held constant, a value is reached where the stage oscillates. Another cause of feedback is the common coupling of the plate and grid circuits through the common current that passes through the cathode. Even with the cathode pin grounded directly, the cathode impedance is far from negligible at high frequencies. This effect is reduced if the tube used has two cathode connections. The plate is returned through its bypass capacitor to the grounded side of the cathode connection, with the grid returned to the ungrounded side, and the paths of the two currents are comparatively independent. When the cathode is above ground, because of impedance in a cathode bypass capacitor, there is a further danger of instability. This is overcome by using capacitors which contain enough self-inductance to make them resonate at the intermediate frequency. In this way they present a very-low impedance path to ground. High-gain i-f stages usually do not use cathode bias.

(4) When the cathode-to-ground path is capacitive, a negative load is placed on the grid circuit, raising its *Q*. Consequently, its impedance is raised, causing instability. Negative loading can be reduced by using a cathode circuit that is inductive at the operating frequency. Inductive cathode loading places a positive load across the tuned grid circuit which lowers the gain. For further stability, the tuned circuits always are returned to ground at the cathode of each stage. Otherwise, currents that flow through a ground common to the input and the output circuit could interact inductively, causing feedback. To reduce feedback from stage to stage, the grounds for each stage are made at one point, and the over-all grounding plan for the amplifier is arranged carefully to prevent interaction between high- and low-level ground circuits.

(5) Instability is caused between stages by direct coupling through the common filament supply and through a common plate supply impedance. This type of instability is overcome easily with suitable r-f chokes and bypass capacitors placed in the appropriate leads in each stage. Careful layout of the power-supply leads and shielding also is necessary to prevent inductive and capacitive interaction between successive stages. The leads to the transformers and the associated parts within the individual stage must be kept as short as possible to prevent the development of stray capacitance which could cause feedback energy. The use of seamless drawn cans and iron slugs that magnetically shield the ends of the coils reduce the reaction through magnetic and capacitive coupling between stages. Capacitive coupling between the windings of the transformer itself can be reduced by placing an electrostatic grounded shield between the windings, or by using shielded wire for the primary. This prevents transfer of energy by direct capacitance between the coils and confines it to inductive coupling alone.

b. High-Gain I-F Amplifiers.

(1) Since maximum over-all gain in the i-f amplifier permits the detector circuit to operate at the high level desired, some means of increasing the gain of the amplifier must be devised. Increasing the ratio of inductance to capacitance in the i-f transformers increases the gain, but when the capaci-

tance in the grid circuit is reduced to a low value the Miller effect becomes prominent. Miller effect introduces a change in the input capacitance when the transconductance of the tube changes and if very low values of input capacitance are used, the changes are sufficient to detune the stage. When sharp impulse-noise bursts cause the signal at the grid of the last i-f stage to increase, they momentarily swing it far into the cut-off region. This momentary swing causes detuning of the transformer because the input capacitance can change as much as 2 $\mu\mu$f. The sudden change in tuning spoils the symmetry of the amplifier response and produces phase distortion that appears in the output as noise or garbling of the signal. Most of the amplitude variation that is produced by this impulse-noise is removed in the detector, but the phase distortion remains. Since the gain of an amplifier increases as the product of the transformer capacitances decreases, it is possible to overcome the Miller effect on impulse noise by using the stray capacitance to tune the output circuit, and a fairly high capacitance in the input circuit. The use of the stray capacitance to tune the plate circuit of the i-f stage allows the ratio of the inductance to capacitance to be increased Increasing the grid circuit capacitance causes the capacitance resulting from Miller effect to be effectively in parallel with a large capacitance and reduces its effect on the tuning of the stage

(2) A second problem in high-gain i-f systems is the instability that begins to appear when the over-all gain is high enough to supply the 2 volts needed to operate the detector in many receivers. The maximum gain of the i-f stages cannot produce the required output without feedback when the r-f signal at the antenna terminals is small. This disadvantage can be overcome partially with a double-conversion receiver. The second i-f section, however, does not have sufficient bandwidth to handle a wide-band signal, and, because of difficulty with spurious responses from the second oscillator, it is not always desirable to use a double-conversion receiver.

(3) A circuit has been devised for use with both wide- and narrow-band f-m that overcomes this disadvantage. After passing through three high-gain i-f stages, the signal is applied to a frequency doubler. Since an f-m signal is not distorted by doubling its frequency, the character of the modulated wave is changed in no way. However, it does make possible a second high-gain i-f strip that is at twice the frequency of the first. The second i-f operates at a frequency with twice the deviation and bandwidth of the first i-f section. Therefore, the signal applied to the f-m detector has a higher deviation without any significant increase in noise input. There is no danger of interaction between the high-level stage near the detector with the low-level input, since the frequencies are different. With most signals, there is sufficient voltage at the grid of the doubler to make it draw grid current, which makes the output of the doubler substantially independent of the input signal. The result is a limitation of variations in amplitude and consequently improved performance. Limiter stages which follow can have more gain too, because there is less danger of feedback. Finally, there is automatic squelching of background noise if no signal is present at the grid of the doubler, because there is no i-f output at the higher frequency.

75. Limiter Circuits

a. Purpose. A limiter stage generally is an i-f amplifier so arranged that, after a certain point, a further increase in input signal produces no change in the amplitude of the output signal. If there is sufficient gain ahead of such a stage, variations of amplitude in the received signal are removed. Since the detector responds

only to frequency variations, the intelligence contained in the modulation is not impaired; in fact, fading and some types of noise are removed. Limiter stages are used after the last i-f stage in some f-m receivers. The circuit of a limiter is almost identical with that of a standard i-f amplifier, with the exception of the values of the components and operating potentials. Although a limiter provides some amplification before its saturation point is reached, its main purpose is to limit the amplitude variations caused by fading and noise in the output voltage.

b. Operation.

(1) The response of an ideal limiter is shown in A of figure 136. The amplifier tube must have a sharp cut-off characteristic. Very low values of screen voltage are used, and little bias is applied to the control grid. Therefore, large values of positive grid voltage quickly drive the tube to saturation, and large negative values drive it to cut-off. The transfer characteristic of the limiter is shown in B. An input signal varying greatly in amplitude with sharp pulses of noise superimposed is applied to the input of the limiter system. The varying voltage applied at the grid of the tube drives it to cut-off or saturation so that the large variations in amplitude are clipped off. In B, some amplification takes place for voltages between points 1 and 2, but the output is held constant beyond 2. Values more negative than point 1 are below cut-off and therefore do not appear in the output. For proper limiter action, the lowest-amplitude signal must swing at least from 1 to 2. The resultant output signal is square because the extreme positive and negative peaks have been clipped off in the limiter.

(2) The limiter circuit shown in figure 137 is similar to the standard i-f amplifier

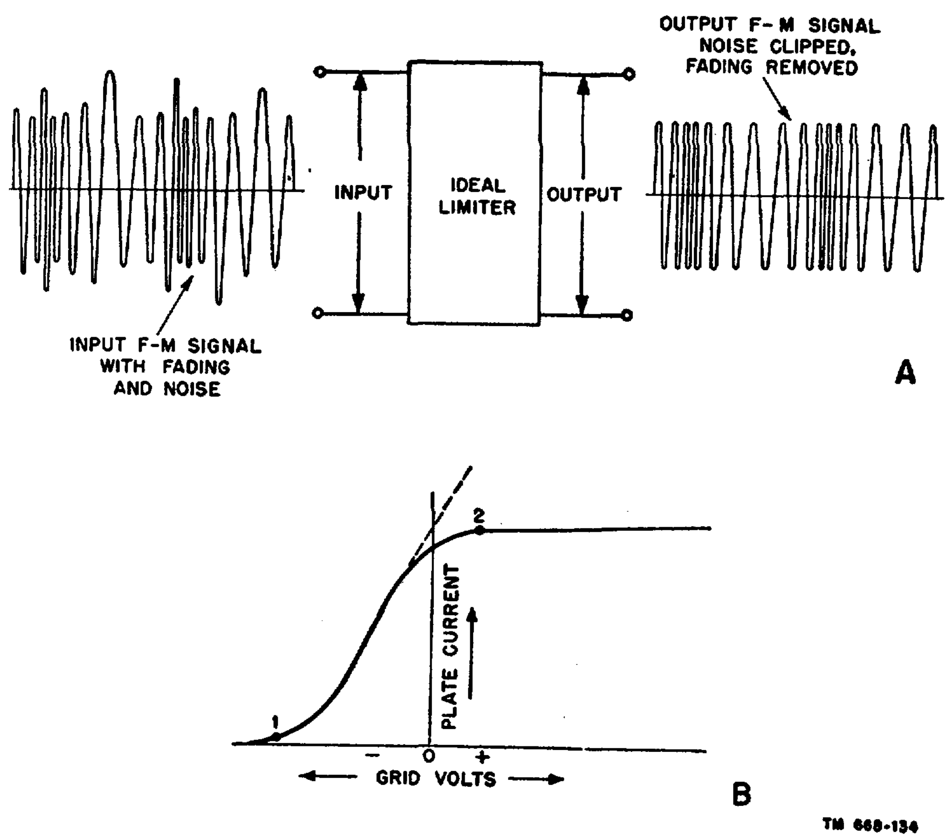


Figure 136. Ideal limiter and its characteristic.

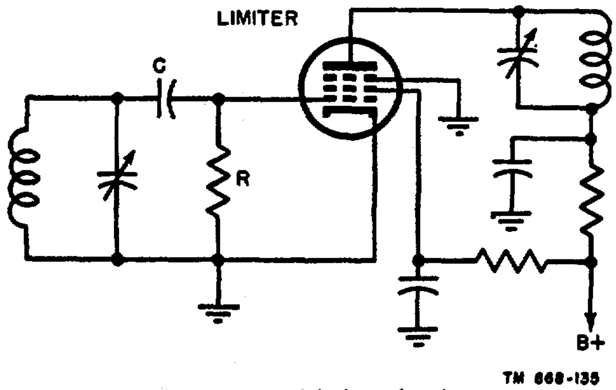


Figure 137. Limiter circuit.

except for the grid-leak resistor and capacitor to provide bias. The R-C combination produces a voltage that is equal to the peak d-c rectified voltage between grid and cathode. The rectification that takes place provides the bias, in addition to its clipping action. The short disturbances in the signal caused by noise voltage that are longer than the time constant of the R-C circuit are clipped off. The longer variations caused by *fading of the* signal appear in the output. To take care of the slower fading, it is common practice to follow the first short-time-constant limiter with a second limiter that has a longer time constant.

c. Cascade Limiters.

(1) Where it is desirable to eliminate both impulse noise and slow *fading in a* limiter, two limiter circuits with different time constants are required. The limiters usually are in cascade and

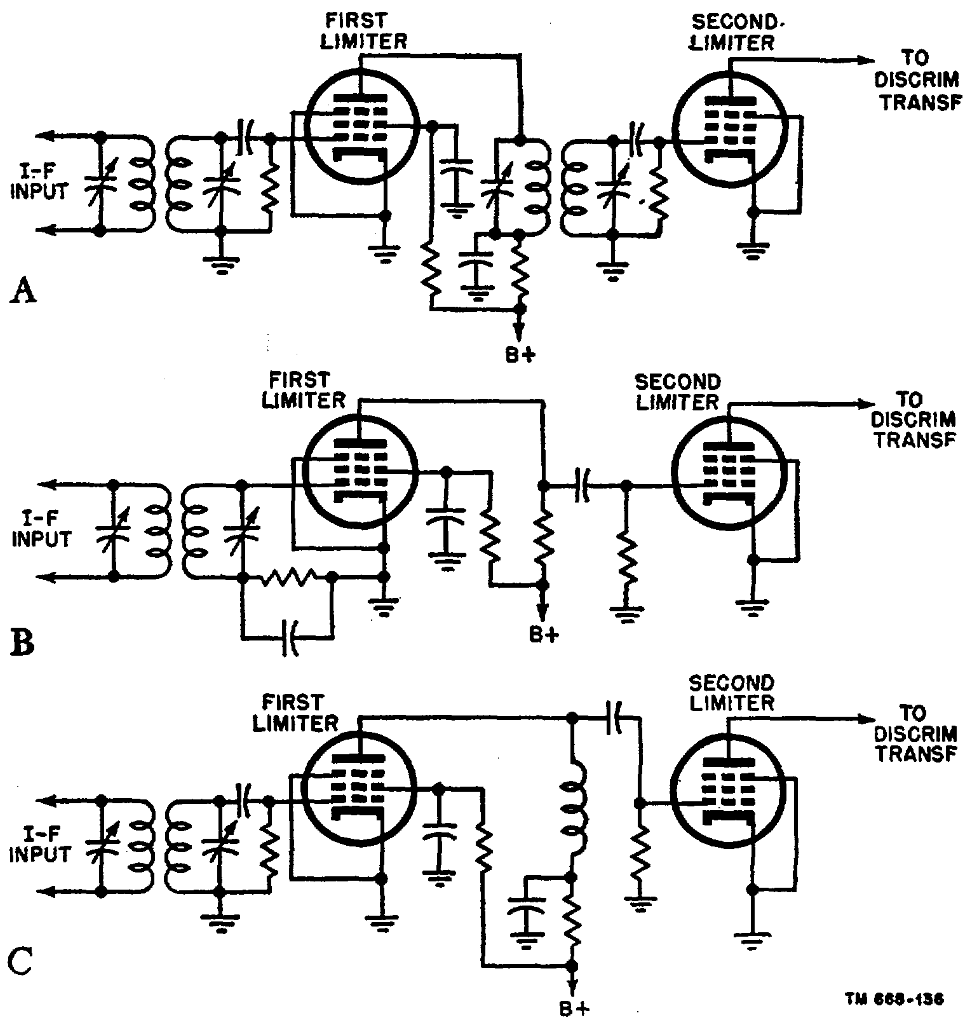


Figure 138. Cascade-limiter coupling circuit.

have different grid-leak values. Several different circuits used to couple such cascade limiters together are shown in figure 138. Transformer coupling, in A, provides greater gain for the second limiter, but is at a disadvantage where there is sufficient i-f gain to saturate the first limiter. Resistance coupling, in B, is most widely used because of its simplicity and ease of adjustment. In C, impedance coupling is a compromise between the other two.

(2) It is not desirable to have any appreciable gain in the limiter because these stages can contribute to the overall amplification of the i-f signal, and may cause regeneration. Amplification in the limiter takes place only when insufficient signal is applied to the grid circuit. Since it takes about 2 volts to saturate the average limiter, if signals of 1 microvolt are to be received at the antenna, a gain of at least 2,000,000 must be effected ahead of the limiter.

d. Miscellaneous Limiters.

(1) Other circuits which serve as limiters, such as saturated diodes and cathode-coupled triodes, are called dynamic limiters. Since the diode ceases to conduct if the plate is made more negative than the cathode, it is possible to clip the negative halves of the signal. If another diode is placed in series with the first, the positive half cycle is clipped also. This circuit produces a square wave from a sine-wave input signal.

(2) A cathode-coupled amplifier also can act as a limiter if proper bias is applied to the grids. As the grid of the first tube goes positive, it draws current. When it goes far negative, the cathode also goes negative, which effectively causes grid current in the grounded grid of the second tube. Output is taken from the plate circuit of the grounded-grid section. The limiting action occurs through grid current on both halves of the cycle. However, since two triodes or a dual triode are required, with little savings in parts, the circuit is infrequently used.

Section VII. F-M DETECTORS

76. Double-Tuned Discriminator

The double-tuned discriminator described in chapter 4 was discussed in conjunction with frequency-control circuits. In f-m receivers the discriminator converts the frequency variation of the signal into audio variations. The transfer characteristic of the discriminator (fig. 139) is plotted in respect to the change in frequency and the amplitude of the audio output voltage. As the frequency of the input varies, each frequency deviation is translated into an effective change in amplitude of voltage at the discriminator output. When the frequency deviation reaches its peak value on either side of the center frequency, the audio voltage also reaches a peak value. Changes in the rate of frequency deviation produce variations in the rate of change of audio voltage which are equivalent to changes in its frequency. The double-tuned circuit is used infrequently in f-m receivers because the three tuned circuits required in the transformer are difficult to aline, and the design of the transformers becomes more critical.

77. Phase Discriminator

a. The phase discriminator, previously discussed in relation to frequency control circuits, is one of the most frequently used f-m detectors. The principal advantage of the phase discriminator over the double-tuned discriminator is its ease of alinement. Since the discriminator has no inherent rejection of amplitude modulation, it always is operated after a limiter stage. The secondary inductance of the discriminator transformer is lower than that of the primary to provide a step-down from the high plate impedance of the limiter to the low input impedance of the diodes. Therefore, the transformer for a phase discriminator differs somewhat from

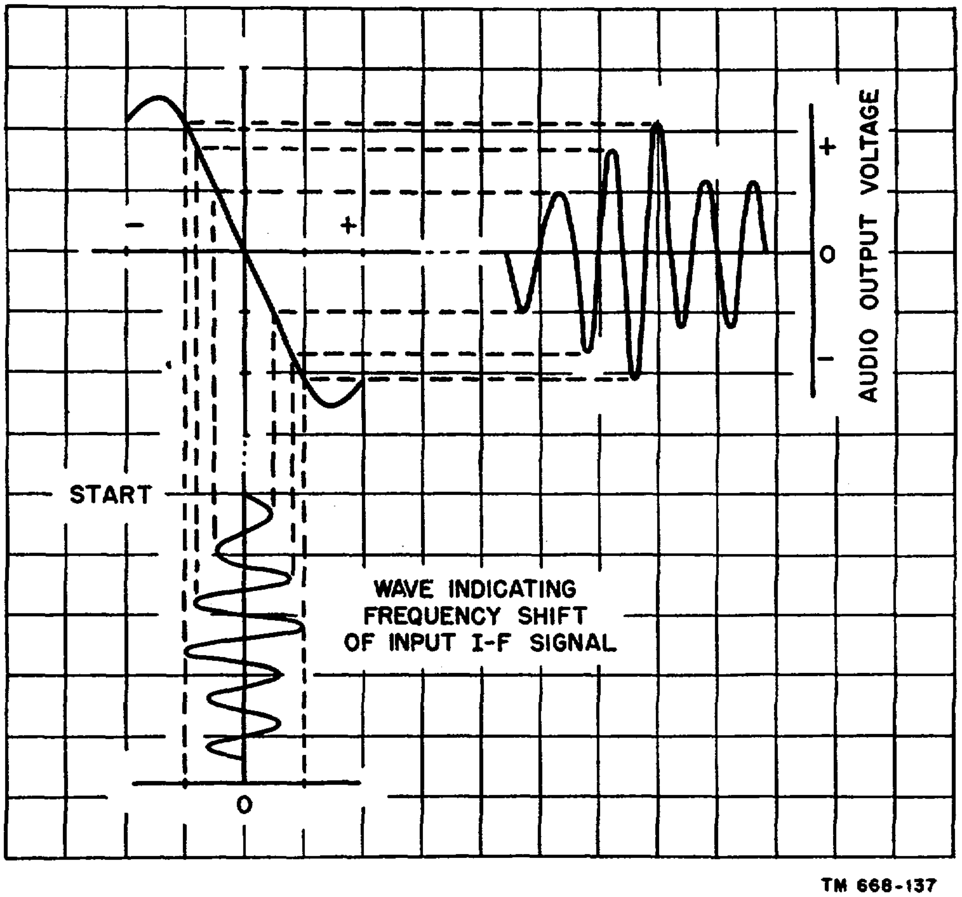


Figure 139. Transfer characteristic of double-tuned discriminator.

that used between two i-f stages. Diodes present a much lower impedance to the load than do the grids of a voltage amplifier, and since they vary from conducting to nonconducting conditions during various parts of the cycle, the actual load presented to the tuned circuit also changes drastically.

b. The modified phase discriminator has essentially the same properties for f-m reception as the ordinary phase discriminator, but fewer components are needed. It is possible to recover recognizable audio signals at frequencies other than the center of the discriminator frequency with any type of discriminator because the selectivity curve of the transformers continues beyond the characteristic S-shaped transfer curve. On the sides of these curves a signal that varies with frequency can be rectified by the diodes. When the receiver is slightly detuned, an audible response is present, although with considerable distortion. This occurs on both sides of the i-f channel, and detection off the center frequency is called side response. With a discriminator, the side responses are sometimes only 20 db lower than the desired channel, which shows the necessity for high adjacent-channel selectivity in the i-f amplifier.

c. A modification of a phase discriminator, designed for operation with a tube that has two diode plates, but only a single cathode, is shown in A of figure 140. The circuit differs from the previous ones in the way the reference voltage is applied to both of the diode plates. Separate windings for each diode are connected by the small capacitor, *C*4. Both windings are tuned by capacitor *C*3, so that the effective resonant circuit is similar to that in other phase discriminators. The i-f voltages are applied across each diode as before, with resistor *R*1 acting as the load for the upper diode; *R*2 acts as the load

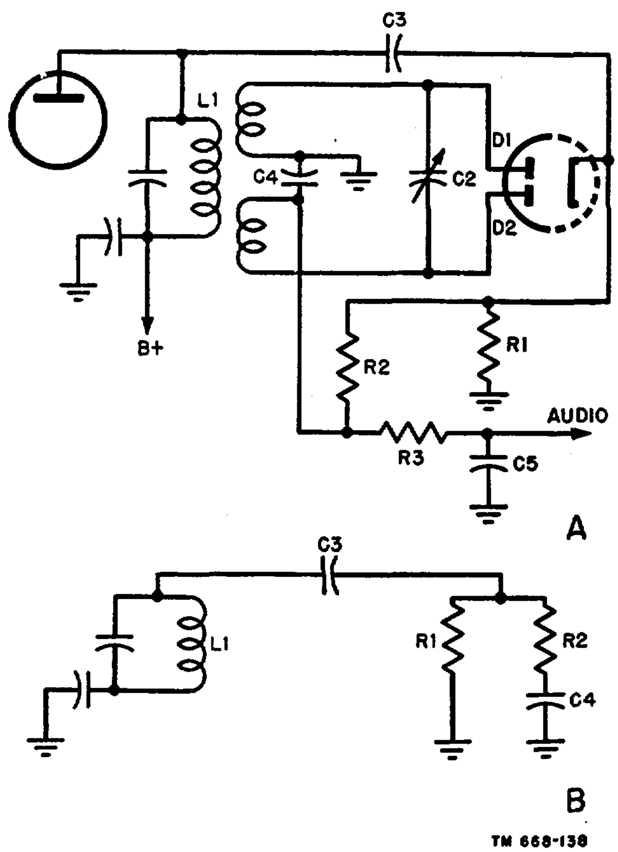


Figure 140. Modified phase discriminator.

for the lower diode because it is returned to the center of the coil and the d-c path is directly from cathode to plate. Although the resistors are connected to a common cathode, they serve as separate loads.

d. The reference voltage is coupled to the cathode by capacitor *C*3, and capacitor *C*4 serves as an r-f bypass to ground for the lower winding. From the i-f stage, there are parallel paths from the plate of the tube, shown in simplified form in B. From the top of the primary, the i-f current passes through *C*3, and then flows through two paths, one through *R*1 to ground, the other through *R*2 and *C*4 to ground. Since the reactance of *C*3 and *C*4 is negligible at the i-f frequency, all of the reference voltage appears across *R*1 and *R*2, as desired. The audio output appears on the high-potential side of the load resistors, which coincides with the lower end of *R*2. The additional resistor and capacitor, *R*3–*C*5, form a de-emphasis network.

78. Ratio Detector

a. Basic Ratio Detector.

(1) The basic purpose of a discriminator is to rectify two i-f voltages whose amplitudes depend directly on frequency. These rectified voltages then are combined so that no voltage appears across their output at the center frequency of the i-f amplifier. A difference voltage proportional to the difference in frequency of the two applied i-f voltages is produced when the frequency of the i-f signal is above or below the center frequency. This detector is insensitive to changes in amplitude at the center frequency, but changes in amplitude off center may cause the audio output to vary. Therefore, whenever a discriminator is used, it must be preceded by a limiter that requires more circuits and tubes. This disadvantage can be overcome by a *ratio detector* circuit which splits the rectified voltages in such a way that their *ratio* is directly proportional to the ratio of the applied i-f voltages, which vary with frequency.

(2) When the sum of the rectified voltages from the transformer is maintained at a constant value, the ratio between them must remain constant, and the individual rectified voltages also must be constant. Output, therefore, is independent of amplitude variations in the signal and no limiter is necessary. A simplified ratio-detector circuit (A of fig. 141) shows both diodes connected so that their output adds, instead of subtracting as in the discriminator. Capacitors C_L across the load resistors have a large value of capacitance and are charged by the output voltage of the rectifiers. This tends to make the total voltage across the load constant over the period of the time constant, $R_L C_L$, since a large capacitor across the combined loads maintains an average signal amplitude that is adjusted automatically to the required operating level. The rectified output

must not vary at audio frequency, and the time constant of the capacitor and the load resistors must be great enough to smooth out such changes. This time constant is approximately 2/10 second. The basic phase comparison circuit and the appropriate vector diagram of the ratio detector and the phase discriminator are the same.

b. Practical Ratio Detector.

(1) In the circuit for a practical ratio detector (B of fig. 141) the voltages, $E1$, $E2$, and $E3$, are obtained in the same way as in the modified phase discriminator. Therefore, the applied voltage to the diodes also is the same. The diodes are connected in series, and the current through load resistor R_L is always in the same direction. Consequently, R_L acquires the polarity shown when the current flows from the plate of $D1$ to the cathode of $D2$. When an unmodulated signal is applied to the primary of the transformer, equal and opposite voltages $E2$ and $E3$ are developed across the secondary in respect to the center tap. These voltages are rectified by the diodes, with the output voltage across the load resistor, equal to their sum, or $E2$ plus $E3$, and the large capacitor, C_L, is charged to this constant voltage. The time constant of R_L and C_L is long compared with the lowest audio frequency.

(2) Since the voltage across C_L is constant, the sum of the voltages across $C3$ and

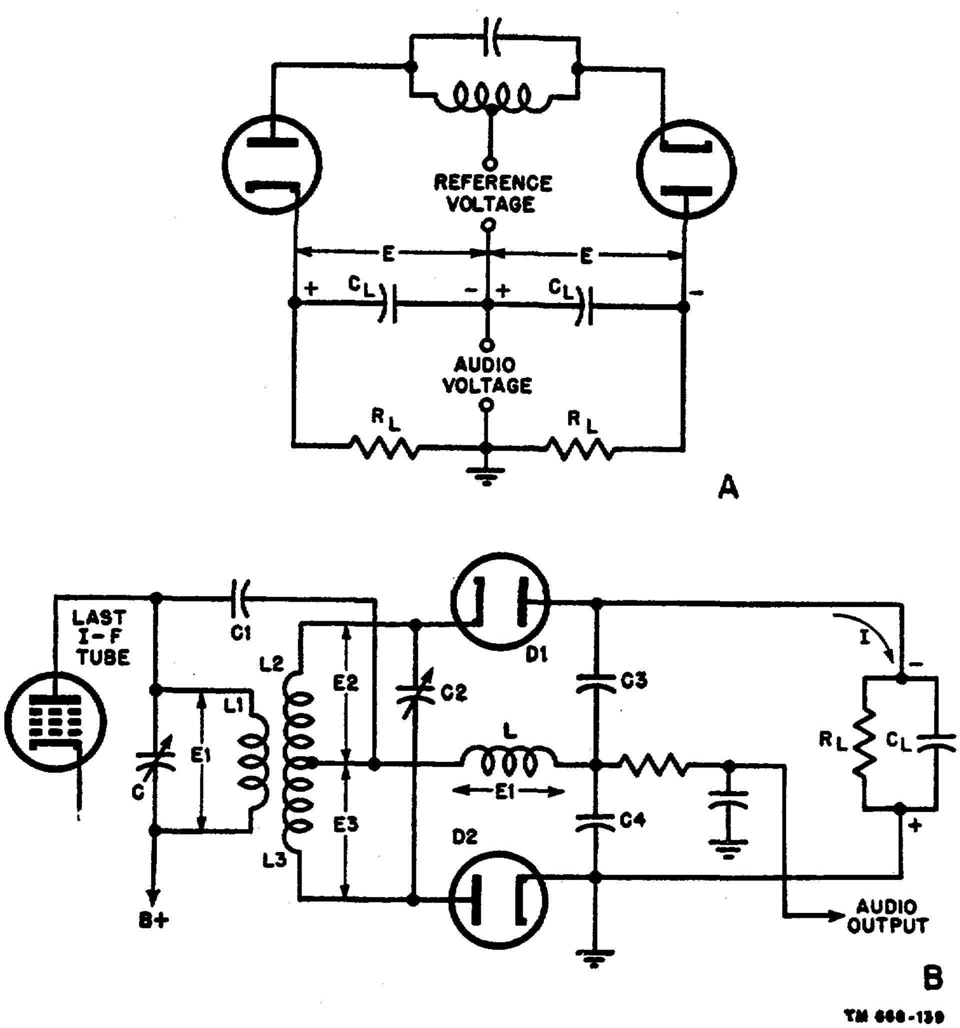


Figure 141. Ratio detector.

$C4$ also must remain fixed. When the carrier frequency shifts with modulation, however, the voltages across $C3$ and $C4$ change, but the sum of their voltages stays fixed at the amplitude of the charge on C_L. When the frequency decreases, $C4$ acquires a greater charge than $C3$; when the frequency increases, $C4$ loses charge to $C3$. Therefore, the voltage between the center tap of the two capacitors and ground varies as the ratio of the voltages across $C3$ and $C4$, the ratio depending on the instantaneous frequency. A variable voltage whose amplitude depends on the frequency deviation of the carrier consequently can be applied to the audio output. As the rate of variation increases with frequency deviation, the voltage at the center tap changes frequency, producing a higher audio frequency. Any amplitude variation in the input signal to the transformers, no matter where the carrier is in its swing, also tends to change the voltage across $C3$ and $C4$. The voltage across the R-C network, however, cannot change rapidly enough to follow the amplitude modulations, and the ratio of the voltage across $C3$ and $C4$ do not change enough to produce an audio output.

c. Performance of Ratio Detector.

(1) The rectified voltage across the load circuit of the ratio detector adjusts itself to the amplitude of the input signal, and there is no minimum level where amplitude variation still can appear in the output. No matter how weak the signal is, the amplitude variations are removed to some extent by the constant charge on the capacitor. However, if signals of greater strength are tuned in, the charge on the capacitor is increased, and the total voltage across C3 and C4 is increased. Consequently, ratio detectors produce audio output that is proportional to the average strength of the received signal. Ratio detectors can operate with as little as 100 millivolts of input, which is much lower than that required for limiter saturation, and less i-f gain consequently is required. This receiver also is relatively quiet when no signal is received, since tube noise is not amplified as much.

(2) As shown by the curves in figure 142, the tuning characteristic of the ratio detector has much lower side responses than the discriminator because they contain appreciable amplitude modulation which is rejected in the load circuit. The disadvantages of the ratio detector are its greater susceptibility to impulse noise and fading, greater

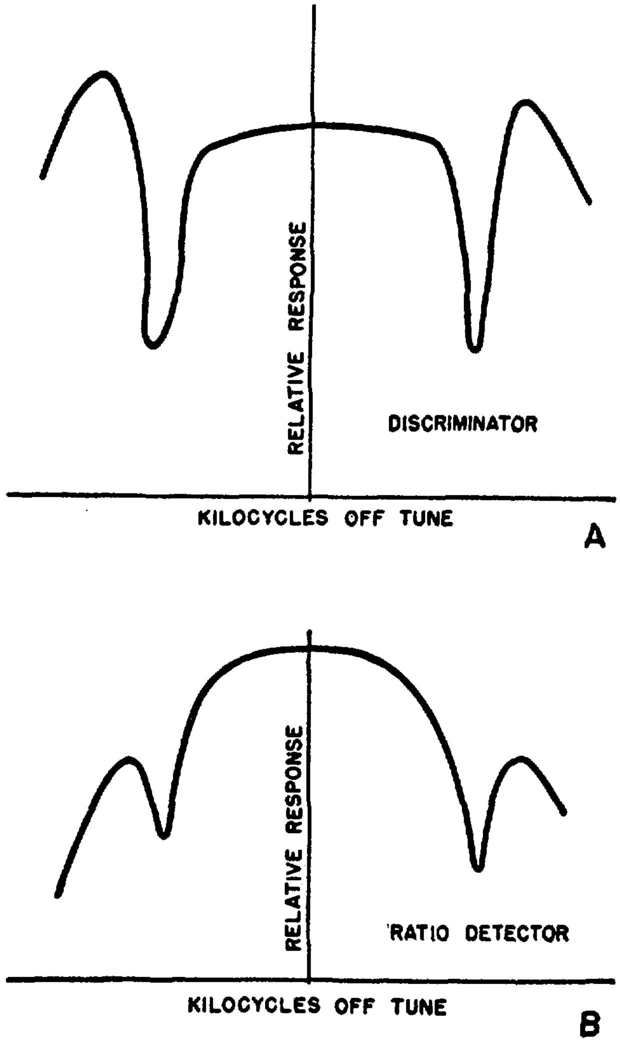


Figure 142. Tuning characteristics of discriminator and ratio detector.

difficulty in alinement, and more complicated transformer design.

d. Modified Ratio-Detector Circuits.

(1) Two modifications of the ratio detector circuit are shown in figure 143. The modification in A allows simpler construction of the transformer. It consists of the addition of a third winding, *L*, from which the reference voltage and the audio output are derived. This tertiary winding consists of a few turns closely coupled to the lower end of the primary. It is essentially a low-impedance source since it is untuned, and the voltage induced in it from the primary is 180° out of phase with $E1$. Voltages $E2$ and $E3$ are each 90° out of phase with $E1$, and the same amount out of phase with the reference voltage which appears across *L*. Capacitor $C5$ is an r-f bypass effectively grounding *L* at the intermediate frequency. Capacitors $C2$ and $C3$ function as ratio capacitors.

(2) The net diode current flow through *L* and $C5$ is zero when the diode currents are equal, because the currents flow through *L* and $C5$ in opposite directions when the carrier is at the resonant frequency of the secondary. When the frequency shifts, the diode currents are unbalanced, producing an audio voltage across $C5$, because it has a high reactance at audio frequencies. This modification overcomes the difficulties of a low-impedance source requirement for the reference voltage, $E1$. At the same time, since it permits the use of a high-impedance primary in the plate circuit of the last i-f transformer, high gain is obtained.

(3) The ratio detector in B has no ratio capacitors, and the cathode of the lower diode has been grounded. The voltage across the tertiary winding is applied to both of the diodes to produce a reference voltage. The i-f current now passes through the large capacitor, $C4$, since the reactance is low and it does not impede the flow of the i-f current. The operation of this circuit is similar to that of the modification shown in A.

79. Synchronized-Oscillator Detector

a. Frequency-Dividing Locked Oscillator.

(1) Using a pentagrid converter tube, an oscillator can be constructed which operates at a frequency that is a fraction of the receiver i-f. If the output of the i-f is applied to what is normally the oscillator grid, with oscillations taking place at the signal grid, the oscillating converter tube will follow the frequency deviation of the i-f at a fraction of the frequency. The oscillator at a lower frequency follows the deviation of the i-f, and the deviation of the locked oscillator is reduced. This reduced deviation can be detected in a conventional discriminator to recover the audio signal.

(2) The oscillator synchronizes with a voltage that is 1/20 of the output oscillation amplitude. This provides an effective voltage gain of 20 on a lower frequency which does not endanger the stability of the i-f amplifier. Moreover, the input to the oscillator has no effect on the amplitude of oscillation. The output is like a limiter in that amplitude variations are removed. Since the discriminator operates on a much lower frequency, the advantages of the high selectivity obtained with double conversion are realized without additional i-f amplifiers or converter circuits. This circuit also discriminates to some extent against impulse noise. Noise immunity is about that of a cascade limiter.

(3) The disadvantages of this system result from the imperfect tuning and insufficient signal strength. Unless the receiver is perfectly tuned, the oscillator does not follow the extreme deviation limits, and distortion is produced in the output. When the signal strength is not sufficient to synchronize the oscillator, no recognizable output at all can be obtained. A more

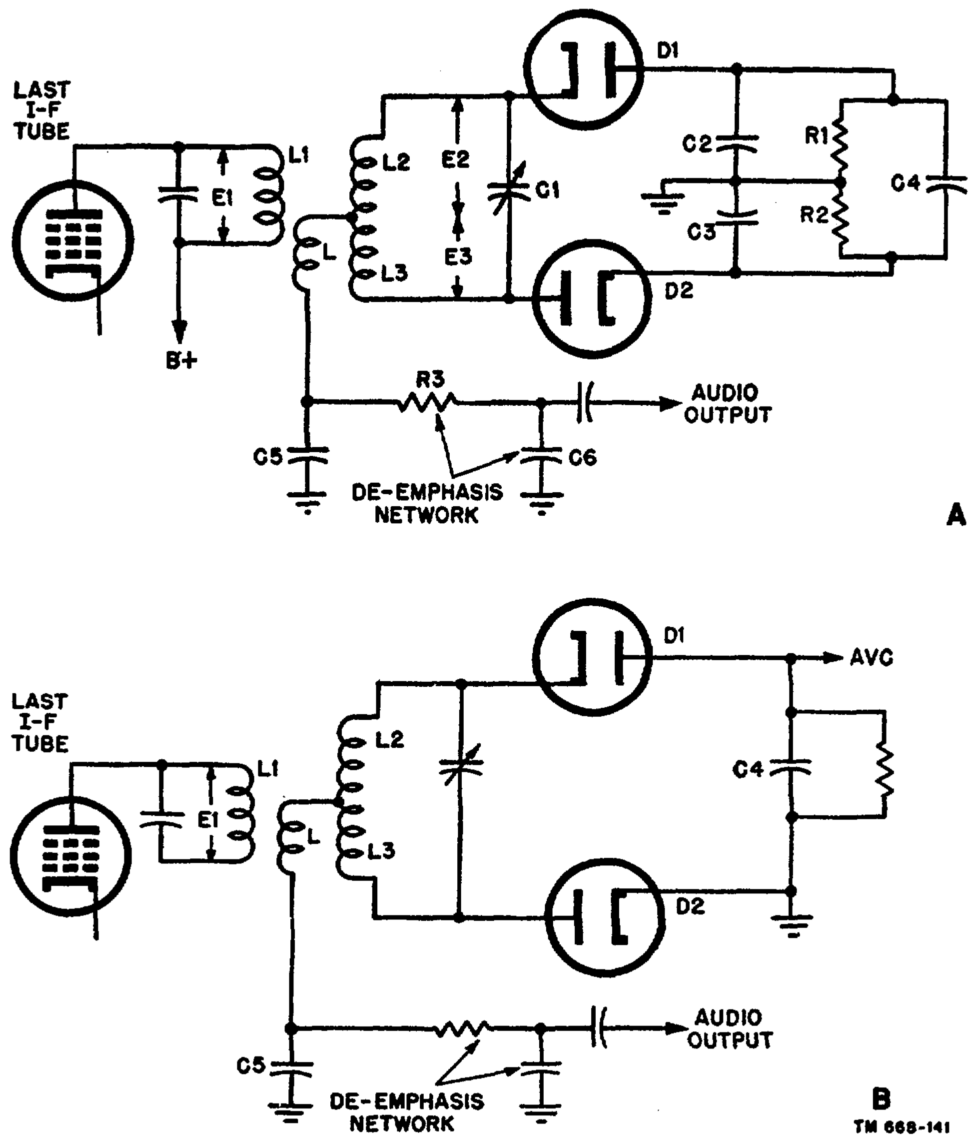


Figure 143. Two modifications of ratio detector.

complicated modification that involves a reactance modulator to keep the oscillator synchronized also has been tried, but the complexity of the resulting circuit makes it impractical.

(4) A major advantage of the frequency dividing locked-oscillator circuit is its rejection of undesired signals in the desired channel. In a conventional f-m receiver, if two stations are on the same frequency but are different in amplitude by 3 db, the stronger signal appears about 9 db greater in the output. With a locked-oscillator detector, a signal that is 3 db stronger than another on the same channel can capture the oscillator almost completely, reducing the level of interference by more than 30 db. This is a great improvement over the other types of detectors.

b. Single-Stage Locked Oscillator.

(1) The frequency-dividing locked oscillator is really not a detector since the actual detection is carried out in a modified discriminator circuit. A special circuit using a tube that incorpo-

rates the detector and the oscillator in a single envelope is illustrated in A of figure 144. The cathode and the first two grids constitute the tube for the locked oscillator, which is a Colpitts circuit. The signal at the oscillator frequency is coupled to the plate of the tube through the electron stream. The oscillator tank circuit is formed by the variable capacitor, *L2*, C_a, and C_b. Grid-leak bias is provided by *C4* and *R1*. The second grid acts as the anode of the oscillator section and is at ground potential for i-f, since it is bypassed by capacitor C_c. The fourth grid is connected internally to the second grid and acts as a shield for the signal grid, number 3, since it is also at ground potential.

(2) The plate circuit is formed by *L3* and *C3* in parallel. All of the tuned circuits are resonant to the i-f. The *Q* of the plate tank circuit is lowered by resistor *R3* in parallel with the coil and therefore its bandwidth is increased. The impedance of this circuit does not change appreciably throughout the i-f pass band. With no signal applied to grid 3, oscillator pulses flow through the plate tank circuit and develop a voltage across it. When a voltage is applied to the signal grid, the amount of plate-current flow changes, and the average value of the pulses of current appearing in the plate load also varies.

(3) A vector diagram of the operation of the stage is shown in figure 144. Vector e_1 is the oscillator voltage on grid 1. Vector e_{3a} is the voltage on grid 3 when the incoming frequency is the center of the i-f band. When there is no signal input, or when the signal is unmodulated, the pulses of current flowing in the tube do not change.

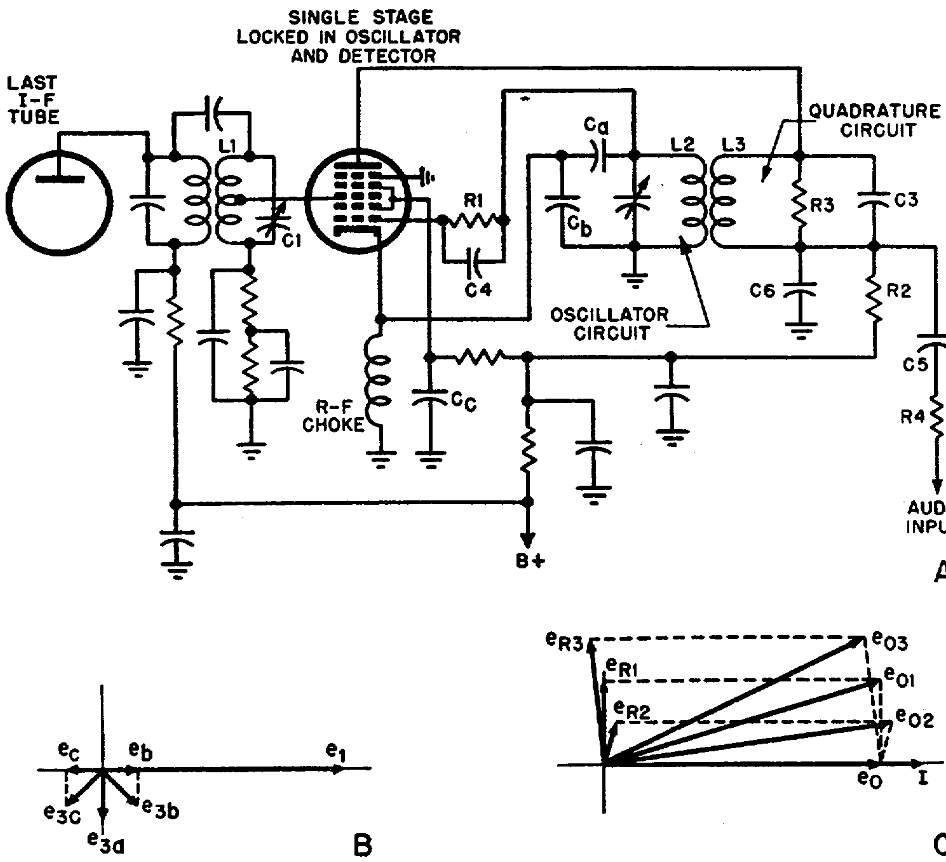


Figure 144. Locked-oscillator detector.

When the incoming signal and the pulses differ in phase, a component of the input signal is either in phase or out of phase with the pulses. The amplitude of the pulse voltage increases or decreases. This phase change is produced when the input voltage frequency varies. Vector e_{8b} shows the signal voltage when the frequency is below the center, and e_{8c} when it is above. These variations produce changes in the main vector, e_1, as shown by the horizontal components, e_b and e_c, that add or subtract from it. The magnitude of the current flowing in the plate circuit therefore varies with the frequency of the incoming signal. The average value of the pulses is obtained in the filter circuit of resistors and capacitors connected to the audio output.

(4) The frequency of the oscillator circuit is changed by feedback coupled from the output circuit. Feedback voltage is proportional to the amplitude of current pulses and is 90° out of phase with oscillator voltage because of the action of the tuned circuit. The circuit acts like an inductively coupled reactance modulator which varies the effective inductance in the oscillator tank circuit. The oscillator therefore tends to stay locked in with the frequency of the incoming signal, as in C. Vector e_O is the voltage across the oscillator tank without feedback. The current flowing through the tank is in phase with the applied voltage, as shown by the current vector, I.

(5) With feedback, the voltage across the tank is the voltage induced by coupling and the vector, e_O. The induced voltage is 90° out of phase with e_O, and this is equivalent to introducing inductance in series with the tank circuit coil. This increased inductance synchronizes the oscillator with the frequency of the incoming signal. With the oscillator exactly synchronized and with no modulation, a voltage is introduced in the oscillator circuit, e_{R1}. With e_O, this produces the resultant vector, e_{O1}. When the incoming frequency increases, the pulses of plate current decrease, and the voltage induced in the oscillator tank also decreases. This is shown by the smaller vector, e_{R2}. The effective inductance is decreased, and the frequency of the oscillator is increased until it is brought back into synchronism with the input signal. When the frequency of the input signal decreases, the reverse operation takes place, as shown by vector e_{R3} and resultant oscillator voltage vector e_{O3}. The plate current changes linearly with these frequency variations. At the same time, it holds the oscillator in synchronism with the applied frequency. The plate current variations are recovered through the audio load resistor, *R*2. Capacitor *C*6 serves to bypass i-f current. The audio output is coupled to the following stages through *C*5 and *R*4.

(6) This circuit effectively suppresses amplitude modulation in the same way that the locked oscillator divider does. In addition, it acts automatically as its own detector. Like all locked oscillator detectors, it has a threshold amplitude of input signal below which no signal at all is received. The input signal required to lock the oscillator is approximately 1 volt, which is slightly less than that needed for a limiter. Amplitude-modulation disturbances appear in the output as distortion because the oscillator is forced out of synchronism for short periods of time, with resultant loss of linearity in plate-current change. This circuit combines the function of limiter and detector in one circuit, with good noise immunity. The audio output is independent of the input signal, although the distortion decreases with increasing signal input. The receiver is quiet in the absence of a received carrier because there is no direct connection between the detector circuit and the i-f amplifier, provided the shielding is adequate.

80. Cycle-Counting Detector

a. An f-m detector that requires no alinement at all, and which operates with a resistance-coupled second i-f amplifier, is shown in figure 145. Selectivity is obtained through the use of a tuned first i-f at high frequency. The resistance-coupled circuit will pass a band of frequencies from low audio frequencies up to about 200 kc. The essential operating feature of this detector is its response to the number of cycles provided by the two-stage resistance-coupled limiter. These limiters produce square waves which are rectified by the dual-diode, whose output consists of negative-going pulses. The positive halves of the square waves are shorted to ground by the first half of the diode. The second half of the diode passes the negative halves, which charge the capacitor connected from plate to ground. As the frequency increases, the average rate of charge increases, and the voltage across the capacitor increases. Starting at a given center frequency, the charge on the capacitor fluctuates, depending on the departure of the signal from that center frequency. The resistor serves to lower the time constant so that the storage of charge can change fast enough to reproduce an audio signal.

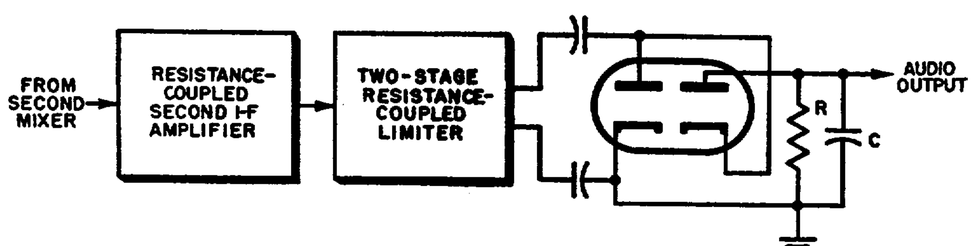


Figure 145. Cycle-counting detector.

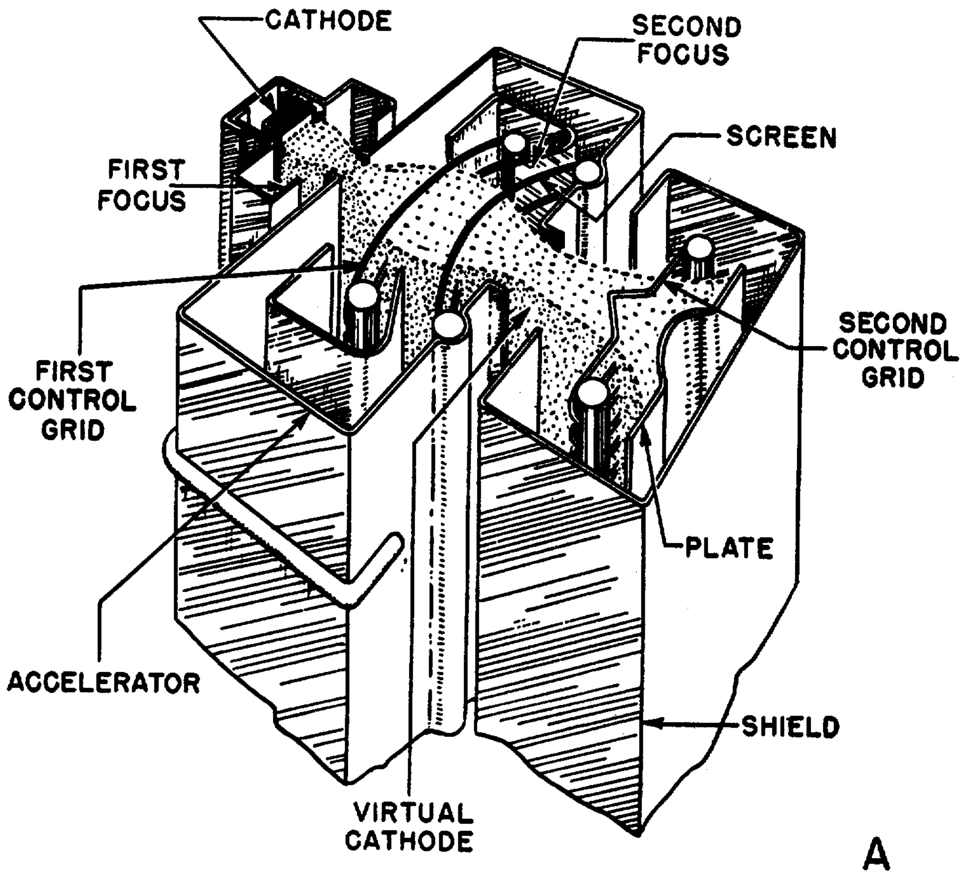
b. The dual limiters reduce undesirable a-m, and the detector responds only to frequency-modulated signals. The output is coupled through a capacitor to the first audio stage. Besides the ease of alinement, this circuit has fairly good sensitivity because of the dual i-f system. The selectivity, however, tends to be poor because of the small number of tuned circuits in the high-frequency i-f. The circuit has low distortion and is used widely in frequency monitors for f-m transmitters. The detector produces output as low as 0 cycle (d-c), which can be used to actuate meters that show the departure of the master oscillator from the assigned frequency.

81. Gated-Beam-Tube Detector

a. Tube.

(1) The gated-beam tube (fig. 146) is a highly efficient limiter and a special type of discriminator contained within one envelope. Its operation depends on the behavior of focused streams of electrons, and, although it has grids and a plate, its operation is different from that of most receiving tubes. With the construction shown in A an unusually sharp cut-off tube results and the transition between anode current flow and cut-off, as controlled by grid voltage, is very abrupt. At the point where current begins to flow, the transconductance is much higher than for any other type of tube. This results in the step-shaped control characteristic shown in B, where plate current is plotted against grid 1 voltage. After the plate current rises to its maximum value, no change in grid voltage can produce any further change in plate current. *This control characteristic is found only in this tube.*

(2) The electron stream from the cathode is passed through focusing and accelerating electrodes which form an electron beam. The beam then passes through the first control grid if this grid is either zero or positive. The electrons continue through a second focusing electrode, a screen grid, and then through a narrow slit which acts as a virtual cathode for the second



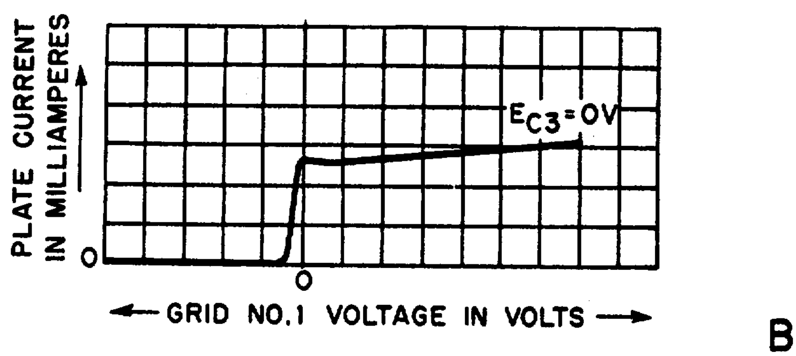


Figure 146. Gated-beam tube.

control grid. The control characteristic of the second control grid is similar to that of the first grid and the screen grid is inserted between them to act as a shield. Electrons pass through the tube in a flat sheet which narrows down at the focus and accelerator electrodes and broadens out in the vicinity of the control grids. When the potential is close to zero, the electrons in the vicinity of the control grid move slowly, although most of them travel in substantially straight lines when passing through it.

(3) If the voltage on the first control grid is made a trifle more negative, a few electrons are repelled in front of it, rapidly building up a space charge

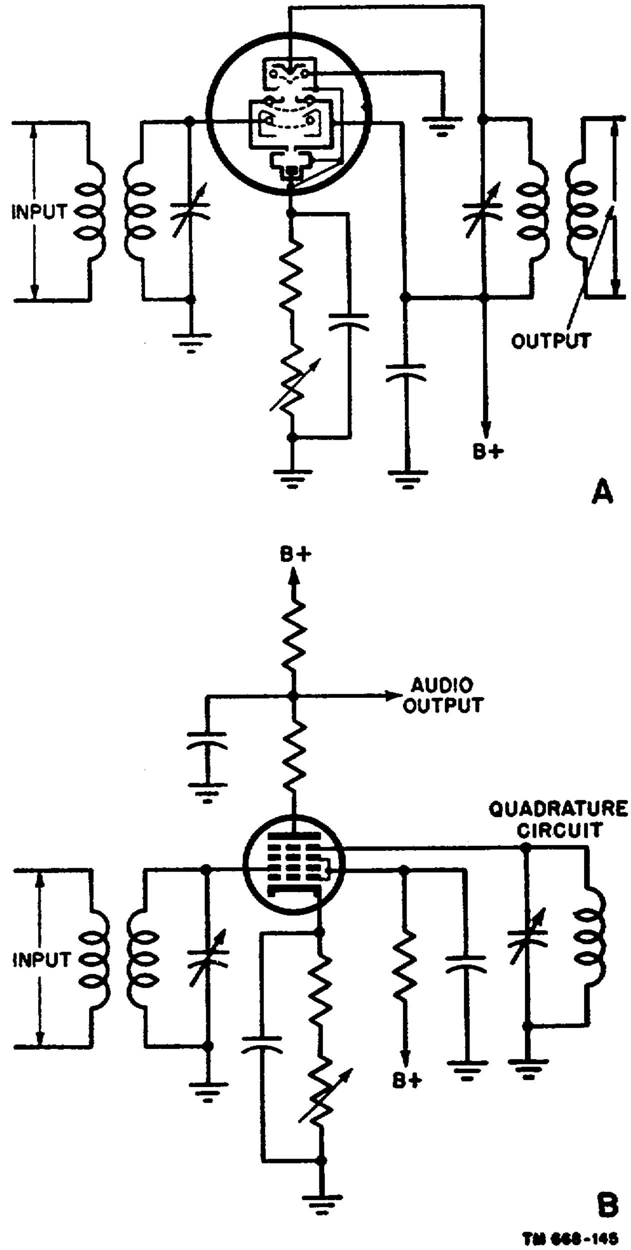
which is concentrated in the center of the beam, forcing the remaining electrons to the side. Those falling back miss the small narrow opening from which they emerged, and the result is an abrupt cut-off of plate current, as shown in B. The operation of the second grid is identical with that of the first except that, since fewer electrons are available, the transconductance is slightly less.

(4) When the control grids are driven positive, they draw current. However, they cannot draw more than a proportionate share of the total beam current. Those electrons not captured by the positive control grid are accelerated and pass through its openings. If the accelerator and anode voltages are sufficiently high, the electrons are moving so rapidly when they near the positive grid that it captures few of them. The result is that the grid current is low, leveling off at a maximum value of about 500 microamperes long before maximum positive grid potential is reached. The entire complex assembly is contained within a seven-pin miniature-tube envelope.

b. Operation as Limiter.

(1) The circuit for the gated-beam tube operated as a limiter (A of fig. 147) is much like that of an ordinary amplifier, but its limiting capabilities are excellent because of the tube-control characteristic. Cathode bias is used, giving an effective negative first control-grid potential of about 1 volt. This corresponds to the center of the steepest part of the control characteristic curve. Since the bias is critical, a small variable resistor is placed in series with the cathode so that field adjustments can be made. The control grid is returned to d-c ground through the secondary of the input transformer. The second control grid usually is grounded and plays no part in the operation of the tube as a limiter.

(2) Limiting occurs instantaneously if the voltage on control grid 1 is shifted



A. Limiter circuit using gated-beam tube.
B. Limiter-discriminator, audio-amplifier circuit using gated-beam tube.

Figure 147. Circuits.

slightly in either positive or negative direction. The application of signal voltage turns the tube plate current on and off with each cycle like a switch. The plate current then consists of positive-going pulses which are flat-topped because the plate current levels off with the application of more than 1 or

2 volts of positive grid signal. Since there are no resistors or capacitors in the grid circuit, this limiter has a short time constant, and impulse noise, which is shorter in duration than a cycle of signal voltage, is clipped almost instantaneously. Consequently, this tube, as a limiter, produces better noise immunity than any other f-m limiter. Operation is extremely simple and stable. Only one stage is needed with far fewer parts than are required for a cascade limiter, whose performance it exceeds.

c. Operation as Discriminator.

(1) When the first control grid is biased at the base of its control characteristic, it passes the beam during positive half-cycles and rejects it during negative half-cycles. The square electron pulses pass through the second accelerator and form a space charge which varies periodically in front of the second control grid. By this space-charge coupling, a periodic charging current of about 15 microamperes per megacycle is produced in the ground-return circuit of the second control grid. If a tuned circuit is inserted in this grid lead, as in B, about 5 volts are built up which lag the input voltage on the first grid by 90°, when the circuit is at resonance. This quadrature voltage operates the second grid on the steep portion of its control characteristic in the same way as the first grid. The result is that a plate current flows which consists of pulses passed by the gating action of the two grids. The beam can reach the plate only when both gates are open, plate current flow starting with the delayed opening of the second grid and ending with the closing of the first.

(2) When the signal to the first grid is frequency-modulated, there is an instantaneous shift in phase between the two grids. The Q of the tuned circuit tends to maintain the frequency, and gating action remains relatively constant on the second grid. The first grid, however, changes in phase as the frequency varies. This of course changes the timing of the opening of the two grids and varies the length of time during which plate current can flow. Consequently, the average plate current varies with the modulating frequency. The plate current pulses charge a capacitor from plate to ground through a small resistance. The audio voltage appears across the capacitor. The typical response curve has a linear relation between deviation and audio output voltage over the entire range of frequencies at which the Q of the quadrature circuit holds up its voltage. This means that a receiver using this tube has a wide range of correct tuning within the i-f pass band and produces little signal outside of this range. The result is better adjacent-channel selectivity than can be obtained with a discriminator or ratio detector.

(3) The performance of the circuit depends on the Q of the quadrature circuit. Higher Q's give greater audio output with a consequent loss of bandwidth. A small resistance in series with the plate circuit causes i-f voltage to appear at the plate. This i-f voltage is coupled through the interelectrode capacitance between plate and second control grid into the quadrature circuit. It is in phase with the quadrature voltage and therefore it aids in driving the tuned circuit. The result is a higher quadrature voltage and greater output.

(4) Effectively, a very linear discriminator and an excellent limiter are combined in one tube, with few parts and with noncritical adjustments. After the limiting characteristic is adjusted by setting the cathode-bias resistor, it is necessary only to tune the quadrature circuit for maximum audio output. High-frequency operation is limited by the capacitance between the signal grid and the quadrature grid. The capaci-

tance is reduced greatly by the screen, but it is sufficient to permit a degenerative, out-of-phase voltage to appear across the quadrature-tuned circuit at high frequencies, thereby reducing its voltage. With great care in shielding, operation can be obtained up to about 30 mc. The sensitivity and a-m rejection decrease as the i-f is increased. Since about 1 volt is needed to operate the tube with full limiting, the same amount of i-f gain is needed as for the pentode grid-bias limiter. However, since the beam tube does not have any gain at the i-f, there is much less danger of instability.

Section VIII. SQUELCH CIRCUITS

82. General

Squelch circuits remove the background noise present in limiter-discriminator detectors when no signal is present. In a high-gain receiver, the noise output is sufficient to be very annoying to operators who must monitor a channel for long periods of time. The squelch circuit lowers the audio output of the receiver when no signal is being received, and allows it to operate normally when a signal is present.

83. Limiter-Derived Squelch

a. Direct-Coupled Cascade Limiter.

(1) The limiter circuit in A of figure 148 produces a partial squelch action. The signal from the i-f amplifier is clipped on the positive half-cycles by the first tube and on the negative half-cycles by the second. Positive grid bias for saturation limiting in the second stage is supplied from the B-plus circuit through *R*1. A large resistance, *R*2, which gives deliberately poor regulation of plate voltage, is used in series with the plate circuit of *V*2. When a signal is applied, the increase in negative bias caused by the negative-going pulses decreases the plate current. The result is that the voltage at the junction of the resistor and the tank circuit rises sharply.

(2) In the region of low plate voltage with no signal, the transconductance of the tube is low and the plate resistance also is reduced. These combine to reduce the noise output of the discriminator. The grid-cathode impedance of the second tube also is lowered by positive bias, which decreases the noise response of the tuned input circuit. When the signal makes the grid bias negative, the plate current drops, raising the plate voltage and reversing these conditions. The plate voltage change is approximately 50 volts with signal applied.

b. Cascade Limiter Disabling Audio Amplifier. This change in plate voltage can be used to disable the first audio stage. In the circuit shown in B, the control voltage across *R*2 is reduced by the voltage divider, R_a and R_b, and applied to the audio-amplifier grid. The cathode bias of the audio tube is adjusted so that the audio grid is sufficiently negative in respect to the cathode to remain cut off in the absence of signal. When a signal is applied, the voltage on the grid rises and the tube again can conduct. The bypass capacitor, C, is used to remove noise and hum from the audio grid.

c. Limiter Squelch Applied to Discriminator Diodes. The squelch voltage derived from the plate circuit of the second limiter can be applied directly to the diode plates of the discriminator, as in C. A positive voltage is placed on the cathodes by the voltage divider, R_aR_b. The diodes then are cut off because of more positive cathode voltage than plate voltage until a signal is received. With a signal present, the plates go positive in respect to the cathode, and the detector operates normally.

84. Discriminator-Derived Squelch

The circuit of figure 149 shows how squelch voltage can be derived from the discriminator circuit. The center of the load resistors, *R*3 and *R*4, of the discriminator develops a negative

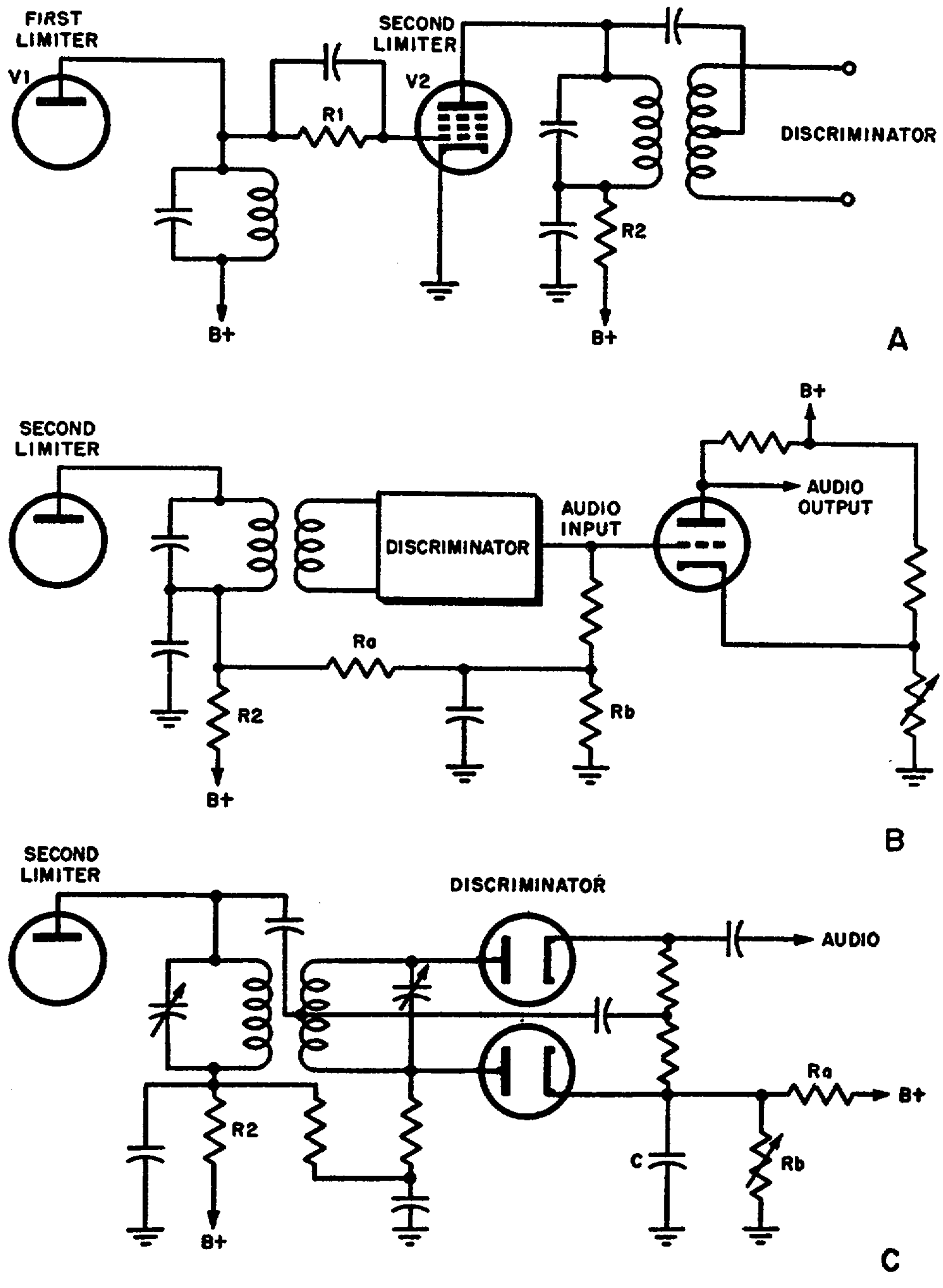


Figure 148. Limiter squelch circuits.

voltage when a signal is applied. This negative voltage applied through *R*5 and *R*6 is sufficient to cut off the squelch tube and permit a signal to pass through the audio amplifier. Without a signal, positive bias, applied to the junction of *R*1 and *R*2, is greater than the bias applied to the cathode of the squelch tube from *R*11 and *R*12. This causes the tube to draw current, dropping the plate voltage and cutting off the audio tube through *R*7.

85. Noise-Rectifier System

a. The circuit of figure 150 shows another type of squelch circuit. With no signal present, a certain amount of noise appears in the output of the discriminator. It is amplified by the noise

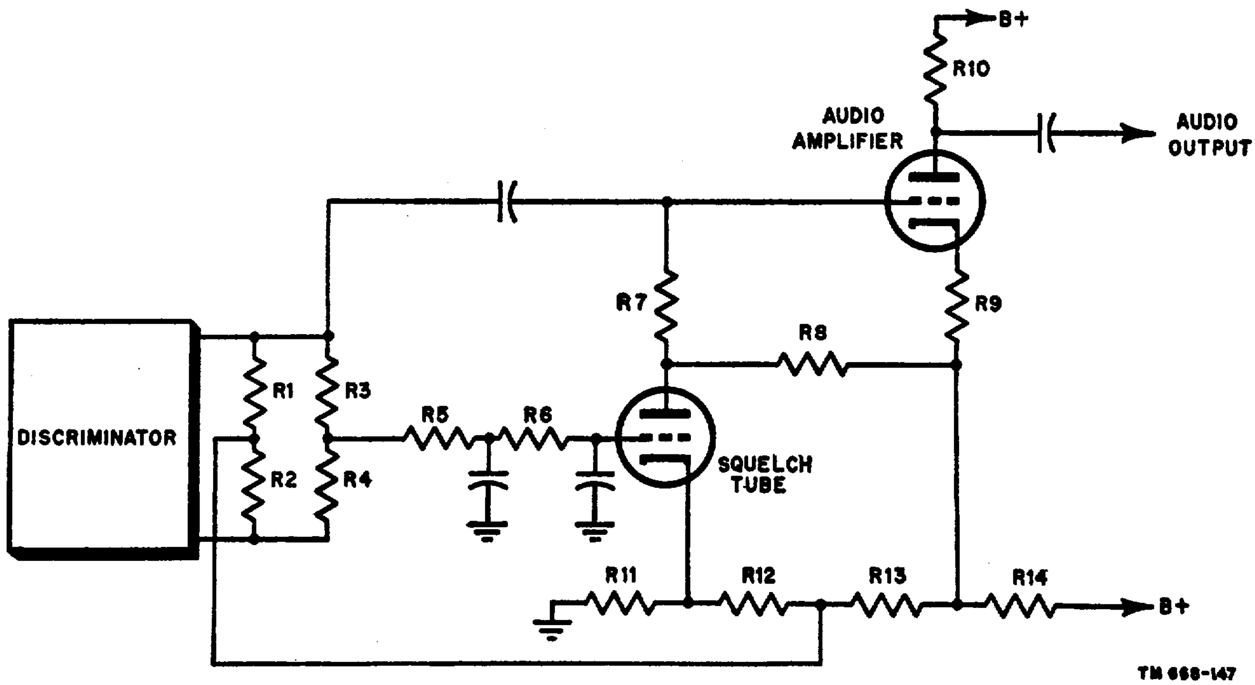


Figure 149. Discriminator-derived squelch.

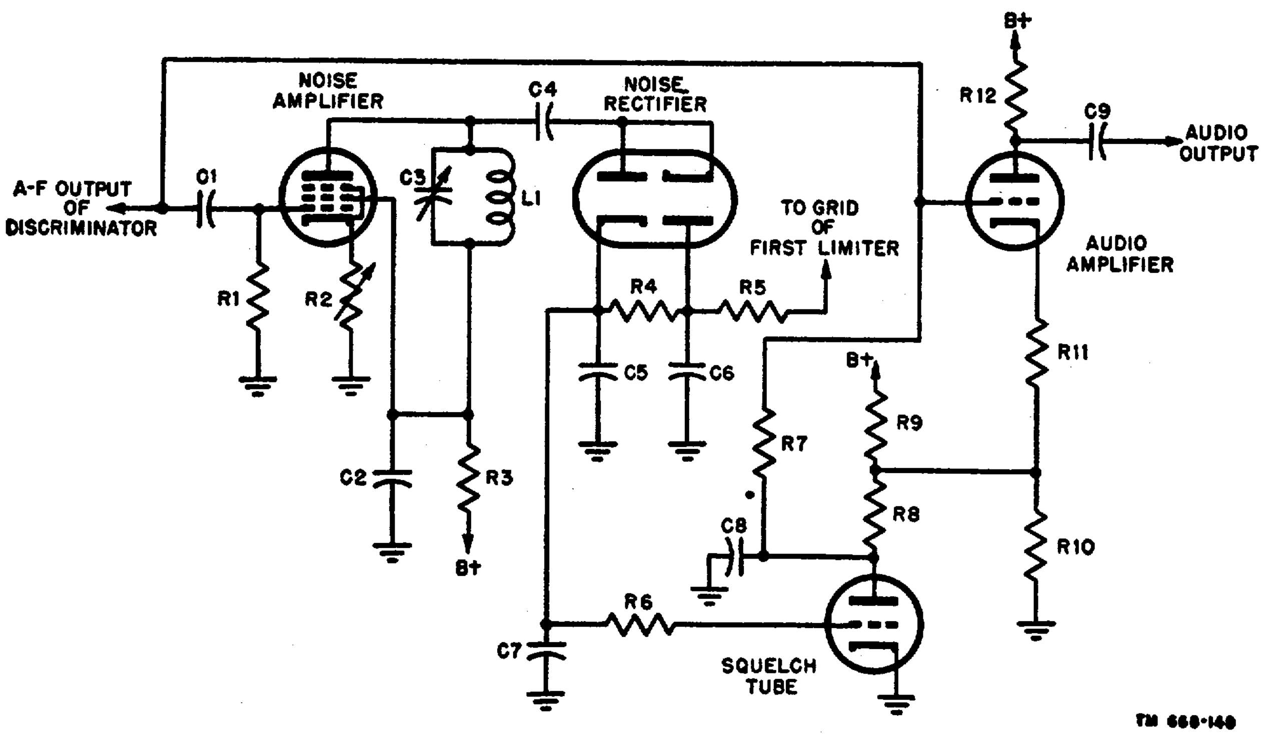


Figure 150. Noise-rectifier system of squelch.

amplifier, and the output is rectified by one section of the noise rectifier. The negative charge built up on *C6* is applied to the grid of the first limiter through *R5*, reducing its gain. The positive peaks are rectified by the other half of the tube, and a positive charge is built up on *C5* and applied to the grid of the squelch tube through *R6*. The squelch tube conducts heavily, cutting off the audio amplifier by making its grid highly negative.

b. A variation of this circuit applies the negative voltage generated by the first limiter grid directly to the squelch amplifier through isolation resistors. A diode is connected across one of these resistors to prevent rectified audio signals in the limiter grid circuit from reaching the squelch tube. In addition, the higher audio frequencies from the output of the discriminator are selected by a high-pass filter network and rectified in another diode and also applied to the squelch grid. With no signal present, a positive voltage is produced by the noise diodes, operating the squelch tube and cutting off the audio amplifier. In the presence of signal, the negative voltage from the first limiter grid cancels this voltage, permitting the audio amplifier to operate. Moreover, the noise-reducing property of f-m causes less positive voltage to be developed at the noise rectifier, increasing the speed of operation of the squelch circuit. This type of squelch can be adjusted to operate on very weak signals.

86. Squelch-Oscillator Circuits

a. The amplifier and voltage arrangement described above are unsatisfactory when the d-c supply voltage is likely to vary over a wide range. In portable equipment powered by batteries, a reliable squelch arrangement is needed which does not depend to so great an extent on voltage variations. The basis for the operation of such a circuit is the abrupt starting characteristic of a pentode oscillator. The circuit diagram of figure 151 shows a typical squelch circuit of this kind. A noise amplifier and rectifier are fed with noise derived from the discriminator output. The rectified noise voltage is applied to a d-c amplifier which is biased positively in the grid circuit to provide an adjustable squelch control. Only the high frequencies are applied to the noise rectifier because it is connected through a high-pass (R-C) circuit that presents a high impedance to any voltages in the speech range. With no signal received, the negative rectified voltage developed at the grid of the d-c amplifier is very large, and the tube is cut off. This produces a high screen voltage, which sends the squelch oscillator-rectifier tube into oscillation. The output is coupled to a second rectifier through a d-c blocking capacitor and is rectified to produce a high negative voltage. When this negative voltage is applied to the grid of the first audio tube, it drives the tube beyond cut-off, eliminating noise from the output.

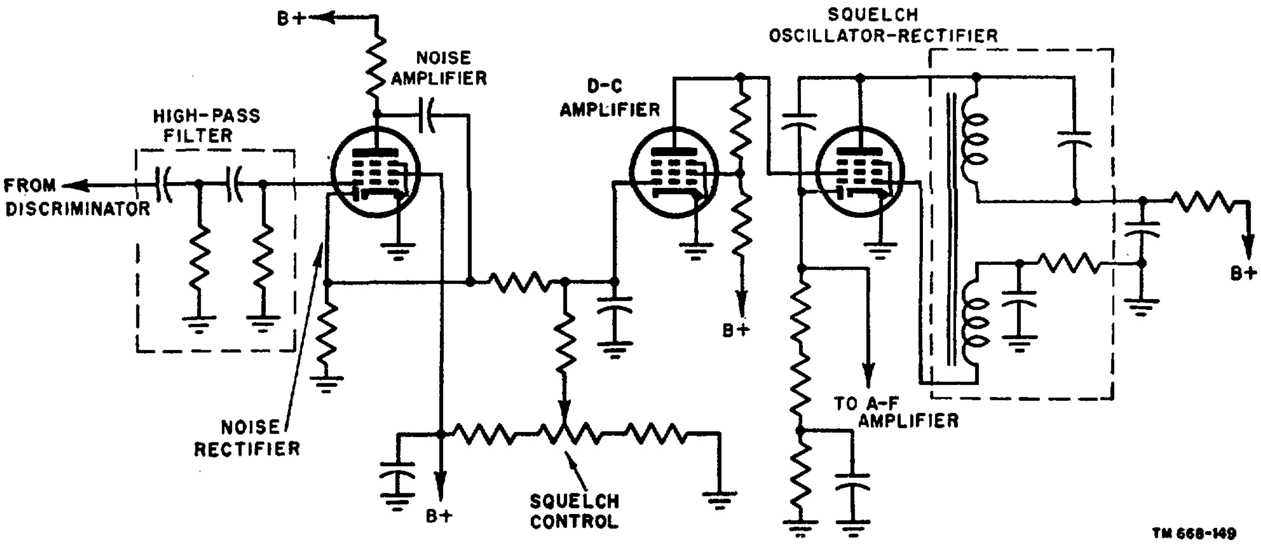


Figure 151. Practical squelch circuit for portable equipment.

b. With signal present, the noise-canceling properties of f-m remove voltage from the noise amplifier and rectifier. The d-c amplifier conducts, dropping the screen voltage, and the circuit no longer oscillates. The rectified voltage from the diode in the oscillator tube falls, and the audio amplifier can pass signals. The point at which the oscillator breaks into oscillation is relatively independent of the plate voltage. As the battery deteriorates, however, the squelch control must be readjusted. The circuit is better than the others that have been discussed, since falling supply voltage does not result in increasing the squelch level.

Section IX. AUDIO AMPLIFIERS, AFC CIRCUITS, AND TYPICAL RECEIVERS

87. Audio Amplifiers

a. The output of the f-m detector is usually about 1 volt. Audio power amplification therefore is necessary to operate loudspeakers, earphones, or other devices. The audio amplifier circuits used in most f-m receivers do not differ materially from their a-m counterparts, except in special equipment. The requirements of adequate response over the entire range of speech frequencies is easily met with simple transformer-coupled class-A amplifiers, using a single-ended beam tetrode or pentode. The inclusion of de-emphasis networks in the audio system is perhaps the only feature that distinguishes an f-m from an a-m audio system. They generally are incorporated between the output of the detector and the grid of the first audio amplifier where they operate at a low-voltage, high-impedance point, so that simple R-C combinations are adequate.

b. Occasionally, where communication is to be carried out under severe conditions, the frequency-response characteristic of the amplifier is shaped so that the lower speech frequencies (below 300 cps) are attenuated and the upper frequencies (above 3,000 cps) are accentuated. This contributes to the intelligibility of speech, since most of the consonant energy is in the upper range. It is possible to achieve a satisfactory result merely by restricting the values of the interstage coupling capacitors so that a rising frequency response is produced. Noise above 4,000 cps then can be attenuated by a small capacitor connected from plate to ground in one of the amplifier stages.

c. In some equipment, a wide a-f channel is required and the frequency response of the amplifier must be flat, from 50 to about 12,000 cps. These amplifiers are sometimes termed *high fidelity*, since a very wide frequency range is reproduced with low distortion. Resistance-coupled and transformer-coupled push-pull audio circuits with high-quality components are used. Inverse feedback, which consists of returning a portion of the output to the input in phase opposition, frequently is used. It results in the reduction of distortion and noise, and the extension of flat frequency response. The equipment which uses this high-fidelity arrangement is found most frequently in large fixed communication centers.

88. AFC Circuits

a. The use of automatic frequency control for receivers was mentioned in chapter 4 in connection with the necessity for keeping transmitters and receivers locked to the same channel. In general, advantage is taken of the discriminator in the receiver by utilizing some of its output to actuate a reactance tube associated with the high-frequency oscillator. This maintains the incoming signal in proper tune, regardless of drift in the receiver or transmitter oscillators. An afc circuit generally is used only when the detector is of the limiter-discriminator type, although it is possible to derive afc voltage from the tertiary coil of a ratio detector.

b. A d-c voltage is produced at the discriminator output if the signal is not exactly in tune with the center frequency. This voltage is positive or negative depending on whether the frequency is low or high. Such a condition takes place if the local oscillator in the front end drifts slightly. This, in turn, changes the relative value of the i-f, and appears as an off-center frequency in the fixed-tuned i-f amplifiers and the discriminator. The d-c voltage produced by the discriminator is applied to the grid of

a reactance modulator, changing the effective transconductance of the tube. The amount of reactance injected into the tuned circuit of the local oscillator then changes, varying the oscillator frequency in such a direction as to cancel out the drift which started the process. The response of the system must be sufficiently slow that variations in carrier frequency caused by modulation are not eliminated. Therefore, a long time constant is necessary in the network that feeds the reactance-modulator grid. In this way, only slow variations in the average frequency operate the reactance tube.

c. Where the transmitter and the receiver are combined in the same unit, the local oscillator of the receiver and the transmitter can be made interlocking and self-stabilizing by the use of various mixers and crystal oscillators. For example, in a typical receiver, the first i-f is approximately 5 mc and the operating frequency is 40 mc. Therefore, if the oscillator is operated on the low side of the received signal, a frequency of 35 mc is needed. The transmitter, by definition, operates on the same frequency as does the receiver. It must, therefore, produce output at 40 mc. A single local oscillator operating at 10 mc can be used for both units in the following way: The fourth harmonic of the oscillator is used directly in the transmitter, and the second harmonic (20 mc) is combined in a suitable mixer with the output from a crystal oscillator operating at 15 mc to give the needed 35-mc local-oscillator frequency.

d. To keep this complicated system interlocked with a distant transmitter requires only that the output of the receiver discriminator actuate a reactance modulator across the local oscillator to maintain the proper 5-mc difference. In doing so, the basic 10-mc frequency is maintained automatically when receiving. In the transmit position the receiver continues to operate, and some of the 40-mc output of the transmitter is received and operates the reactance system, keeping the transmitter locked to the receiver, which in turn already is locked to the distant station. There are other possible combinations of local oscillator, crystal oscillator, i-f, and transmitter frequencies that can be worked out with different arrangements of crystal oscillator and mixers in the same fashion.

89. Typical Receiver Circuits

a. Single-Conversion Superheterodyne.

(1) In the previous sections of this chapter, the operation of the individual circuits that go to make up a complete receiver have been discussed. When the entire receiver is at hand, it is customary to refer to the complete schematic diagram for needed information. At first glance, this diagram may appear complicated but part of the complex appearance can be attributed to the power supply and control circuit wiring, which often is drawn in a confusing manner and basically obscures the real functional circuits. To permit interpretation of a complete schematic, a typical single-conversion receiver is diagrammed in figure 152. The wiring for the filaments, plate supply, and control circuits has been eliminated for clarity. This receiver uses a single r-f stage of the grounded-cathode pentode type, *V1*, a pentode mixer, *V2*, a triode local oscillator, *V3*, three i-f amplifiers, *V4*, *V5*, and *V6*, a ratio detector, *V7*, and two stages of audio amplification, *V8* and *V9*. Although a pentode, such as *V1*, does not result in the most efficient r-f stage, it is simple and stable in operation. The signal from the antenna is introduced by means of a tap on *L*. Capacitor *C1* tunes the circuit to resonance, and is a part of the ganged tuning arrangement that simultaneously tunes the r-f, *C1*, the mixer, *C3*, and the oscillator, *C2*. The output voltage of the r-f stage is developed across *L2*, which is the primary of the transformer. The secondary is the tuned circuit, L3–C3, in the mixer grid circuit and is tuned to the same frequency as *C1*. Trimmer *C27* adjusts the minimum capacitance in the input circuit and partially compensates for differing amounts of reactance that are injected across the input with different transmission lines and antennas.

(2) The input signal to the pentode, *V2*, mixer circuit appears across *L3*. Plate

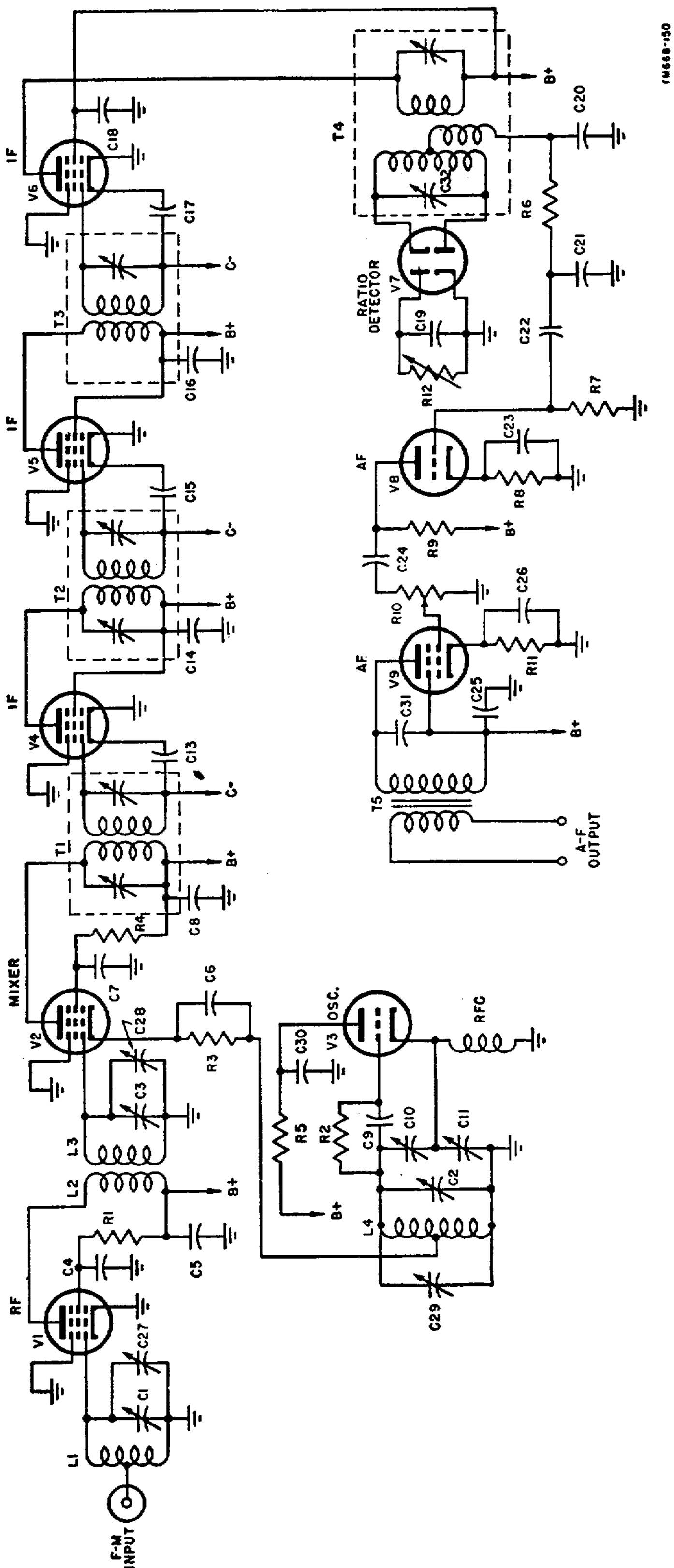


Figure 152. Typical single-conversion f-m receiver.

current is returned to the cathode through a tap on the lower end of the oscillator coil, *L*4. This injects some of the oscillator tank voltage into the cathode circuit through the r-f bypass, *C*6. The output tank circuit of the mixer is tuned to the intermediate frequency and forms the primary of *T*1, which is inclosed in a shielded can. The local oscillator is a Colpitts, with feedback provided by the cathode return to capacitors *C*10 and *C*11. Grid-leak bias is developed across *R*2 and *C*9. The actual tuning is done by the ganged variable capacitor, *C*2, which is across *L*4. The capacitor and resistor network, *C*30–*R*5, in the plate circuit of the oscillator, serves to ground the plate for r-f and prevent oscillator voltage from appearing in the power supply.

(3) The first two i-f amplifiers, *V*4 and *V*5, are sharp cut-off pentodes with separate cathode connections and use fixed bias so that the cathodes can be grounded directly. The grid and plate ground returns are made to the same point on the chassis through *C*13 and *C*14 in the first i-f and through *C*15 and *C*16 in the second i-f. The screen and the plate circuits have a common bypass capacitor, which provides some neutralization of the residual grid-plate capacitance. The primary of transformer *T*3 is untuned, but has sufficient inductance to be resonant at the operating frequency with the various stray capacitances. The capacitor across the secondary lowers the input impedance and reduces the detuning effects of a large signal applied to the grid of *V*6, the third i-f stage. The plate-load circuit of *V*6 is the primary of the ratio-detector transformer, *T*4. The time constant of the stablilizing circuit is adjusted for good a-m rejection by *R*12. Output voltage developed across the tertiary winding of the transformer and *C*20 is transferred to the first audio amplifier through the de-emphasis circuit formed by *R*6, *C*21, and blocking capacitor *C*22.

(4) The first audio stage, *V*8, is a conventional triode, resistance-coupled voltage amplifier with cathode bias furnished by *C*23 and *R*8. The output is developed across load resistor *R*9. The audio signal is coupled to the gain control, *R*10, through *C*24, which is made low in value to introduce some attenuation of the lower audio frequencies for improved intelligibility. The output voltage is developed across the transformer-coupled plate circuit of *V*9. High frequencies, above the useful speech range, are attenuated by *C*31. Actually, this capacitor and the inductance of the primary of *T*5 form a parallel resonant circuit at about 4,000 cps. Therefore, a greater load resistance is presented to the amplifier at this frequency and a greater output voltage is developed across it which serves to accentuate the consonants in the upper speech range. *C*25 serves as both a screen bypass and a plate return for the audio amplifier. Cathode bias is developed across *R*11 and *C*26 for class-A operation, with *C*26 made sufficiently large that the bypass action is poor at the lower audio frequencies. This introduces some out-of-phase voltage on the grid, effectively reducing the gain of the amplifier in this range. The result is a more effective cut-off at the low frequency than is obtained through a low value of *C*24 alone.

b. Typical Double-Conversion Superheterodyne.

(1) The complete schematic of a high-sensitivity, high-stability, double-conversion f-m superheterodyne is shown in figure 153. A direct-coupled, driven, grounded-grid, r-f stage, *V*1 is followed by a triode mixer, *V*2, which is cathode-coupled to a crystal overtone oscillator, *V*3. The first i-f signal is amplified by *V*4 and fed to the second mixer, *V*5, where it beats with the local oscillator signal from *V*6, producing the second i-f. The second i-f is amplified by *V*7 and *V*8 and applied

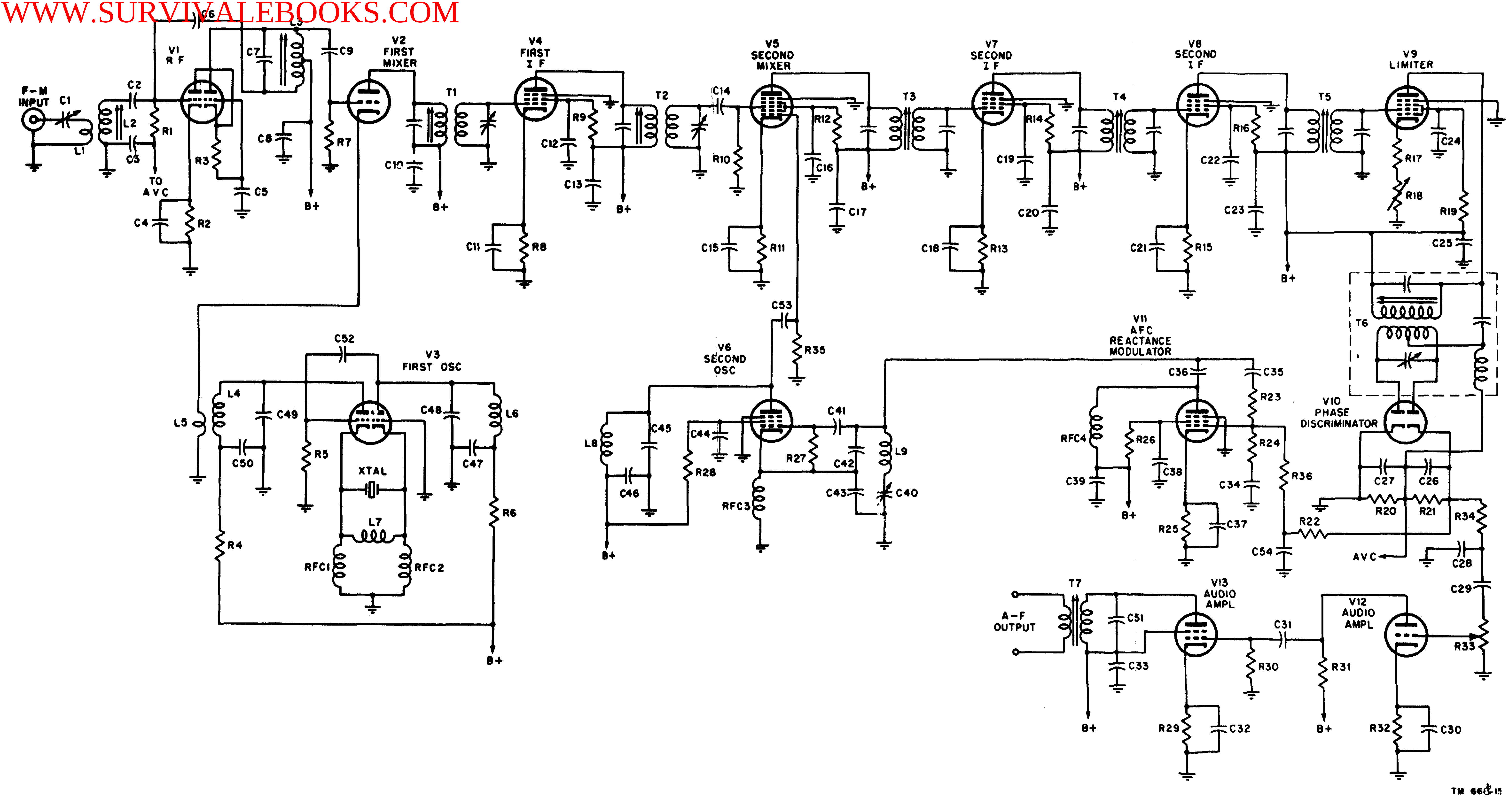


Figure 153. Typical double-conversion f-m receiver.

to the injection grid of a gated-beam tube, $V9$, operating as a limiter. The output of the limiter stage is transformer-coupled to the discriminator, $V10$, from which is derived the audio, afc, and avc voltages. The audio signal is fed to conventional audio amplifiers, $V12$ and $V13$, which are identical to those used in the single-conversion receiver. The afc voltage is applied to reactance tube $V11$, which changes the frequency of the second oscillator to produce a constant second i-f. The avc voltage is returned to the grid of the r-f amplifier.

(2) Since the first local oscillator is crystal-controlled, the input circuits of the r-f amplifier and mixer have a broad pass band. A double-tuned circuit formed by $C1$–$L1$ and $L2$ with stray capacitance is used at the grid of the first section of the driven, grounded-grid circuit. Since avc voltage is applied to the grid, it must be grounded for r-f by capacitor $C3$. The d-c voltage is fed through $R1$, which is isolated from the tuned circuit by $C2$. $C5$ and $R3$ develop the cathode bias for the grounded-grid section of the circuit. The direct-coupled circuit used permits a considerable reduction in the number of parts, and also in stray capacitance. The output signal is developed across the tuned circuit, $L3$–$C7$, which is center-tapped and grounded for i-f by $C8$. To improve the noise figure, the out-of-phase neutralizing voltage at the ungrounded lower end is coupled back into the input circuit by $C6$. Output voltage from the stage is capacitively coupled to the triode mixer, $V2$, through $C9$, and $R7$ serves as the grid-leak bias resistor. The crystal overtone first oscillator is inductively coupled to the cathode of the mixer by the tank circuit of $L5$. The output of the mixer is amplified in the second i-f amplifier, $V4$, and is coupled to grid 3 of the second mixer of the pentagrid converter, $V5$, through the i-f transformer, $T2$. The oscillations in the grid and screen circuits of the second oscillator, $V6$, are coupled to the plate through the electron stream, and appear across the output tank circuit, $L8$–$C45$. The output circuit is tuned to a harmonic of the actual oscillator frequency so that higher stability is obtained through isolation of the frequency-determining tank formed by $L9$, $C42$, $C43$, and $C40$. $R27$ and $C41$ serve as the grid-bias combination, and the r-f choke in the cathode permits the r-f voltage to develop across the feedback capacitors. The output of the second local oscillator is fed to grid 1 of the mixer through the $C53$–$R35$ network. The frequency of the oscillator is controlled by the reactance modulator of the afc circuit, $V11$, which injects inductive reactance into the tank circuit to compensate for any oscillator drift.

(3) Signal voltage from the second mixer is amplified by the conventional tuned i-f amplifiers, $V7$ and $V8$, with the needed selectivity provided by transformers $T3$, $T4$, and $T5$. The output of the secondary of $T5$ is applied to the first grid of a gated-beam-tube limiter, $V9$, which removes amplitude variations and impulse noise from the signal. The limited output is coupled through $T6$ to a conventional phase discriminator, which also produces the afc and avc voltages. Afc voltage is coupled to the grid of the reactance modulator through $R22$. The reactance modulator is prevented from responding to the variations in the signal by $R36$ and $C54$. $R34$ and $C28$ act as a de-emphasis circuit at the output of the discriminator. The remaining audio section of the receiver is the same as in the single-conversion receiver. Analyzing a complicated receiver circuit is a simple procedure when its component circuits are broken down into separate stages. With the inclusion of the control and power supply, the analysis is similar, although it sometimes is more difficult to see the interconnection of circuits and functions of the individual stages.

Section X. TYPICAL F-M RECEIVERS

90. Single-Conversion F-M Receiver

A single-conversion receiver designed for use in vehicular or ground installations is shown in figure 154. Three separate receivers, alike except for the tuning range, cover the band of frequencies between 20 and 55 mc. The single-conversion receiver is small and compact, with all operating controls and cable connectors mounted on the front panel. Power is supplied by a storage battery and vibrator power supply. The stages include an r-f amplifier, a mixer, a local oscillator, four i-f amplifiers, a limiter, a discriminator, and a two-stage audio amplifier. The squelch circuit uses the pentode section of a pentode-diode tube as a tuned-plate tuned-grid oscillator. The rectifier section rectifies the output and applies a bias to the r-f amplifier tube to reduce its gain. It also applies bias to the audio-amplifier circuits, making them inoperative when no signal is applied. An audio output of 1 watt can be supplied for a speaker or 50 mw for phones.

91. Double-Conversion F-M Receiver

The receiver chassis for the 30- to 40-mc band shown in figure 155 is housed in the same cabinet as the indirect f-m transmitter shown in figure 101. This double-conversion receiver uses crystal oscillators for both low- and high-frequency conversion. A single r-f stage feeds the f-m signal to the mixer where it mixes with the signal produced by a crystal oscillator. The mixer output of the i-f is 4.3 mc and this output is amplified by an i-f amplifier and fed to a

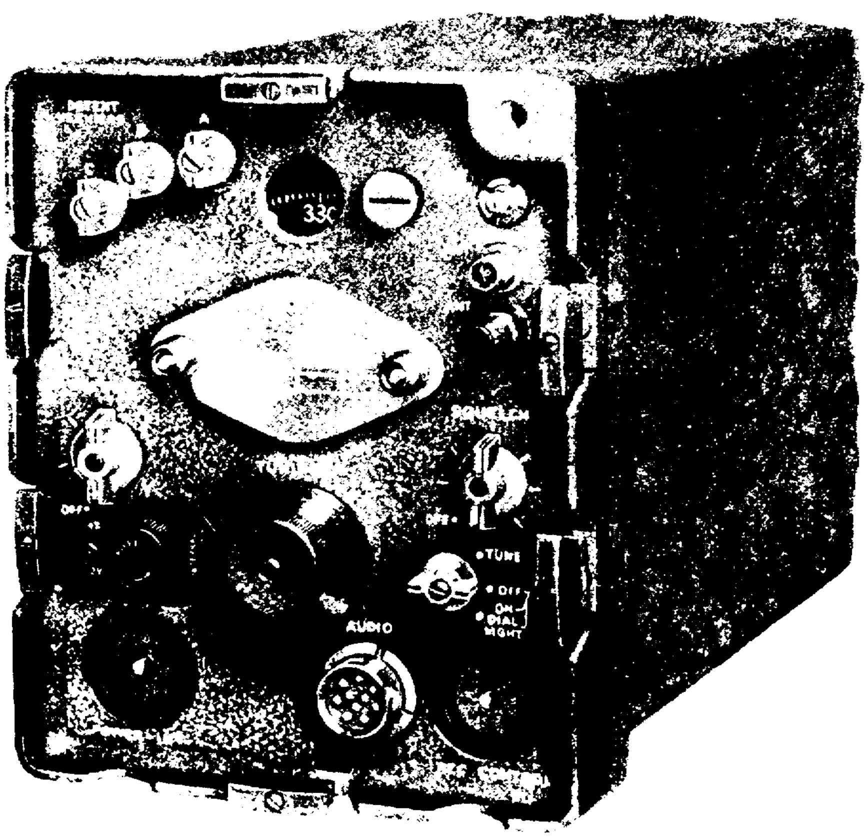


Figure 154. Typical single-conversion receiver.

triode-hexode mixer. The triode section is a crystal oscillator. The output of the hexode mixer section is at the low i-f of 455 kc. The low i-f amplifier feeds the limiter-discriminator section. A cascade limiter feeds the discriminator, which supplies an a-f signal to a two-stage audio amplifier. The output of the discriminator also is used to feed a noise amplifier and rectifier, which in turn inject a signal into the squelch circuit. The squelch tube cuts off the first a-f amplifier when no signal is present. If a signal is present, the first limiter renders the squelch circuit inoperative, allowing the first a-f amplifier to conduct normally. The final a-f amplifier provides sufficient output to drive a loudspeaker.

92. Walkie-Talkie

a. A versatile, battery-powered, f-m receiver-transmitter, designed to provide manpack communications for armored, artillery, and infantry units, is shown in figure 156. It also can be used in airplane and vehicle installations, or in semipermanent ground installations. The walkie-

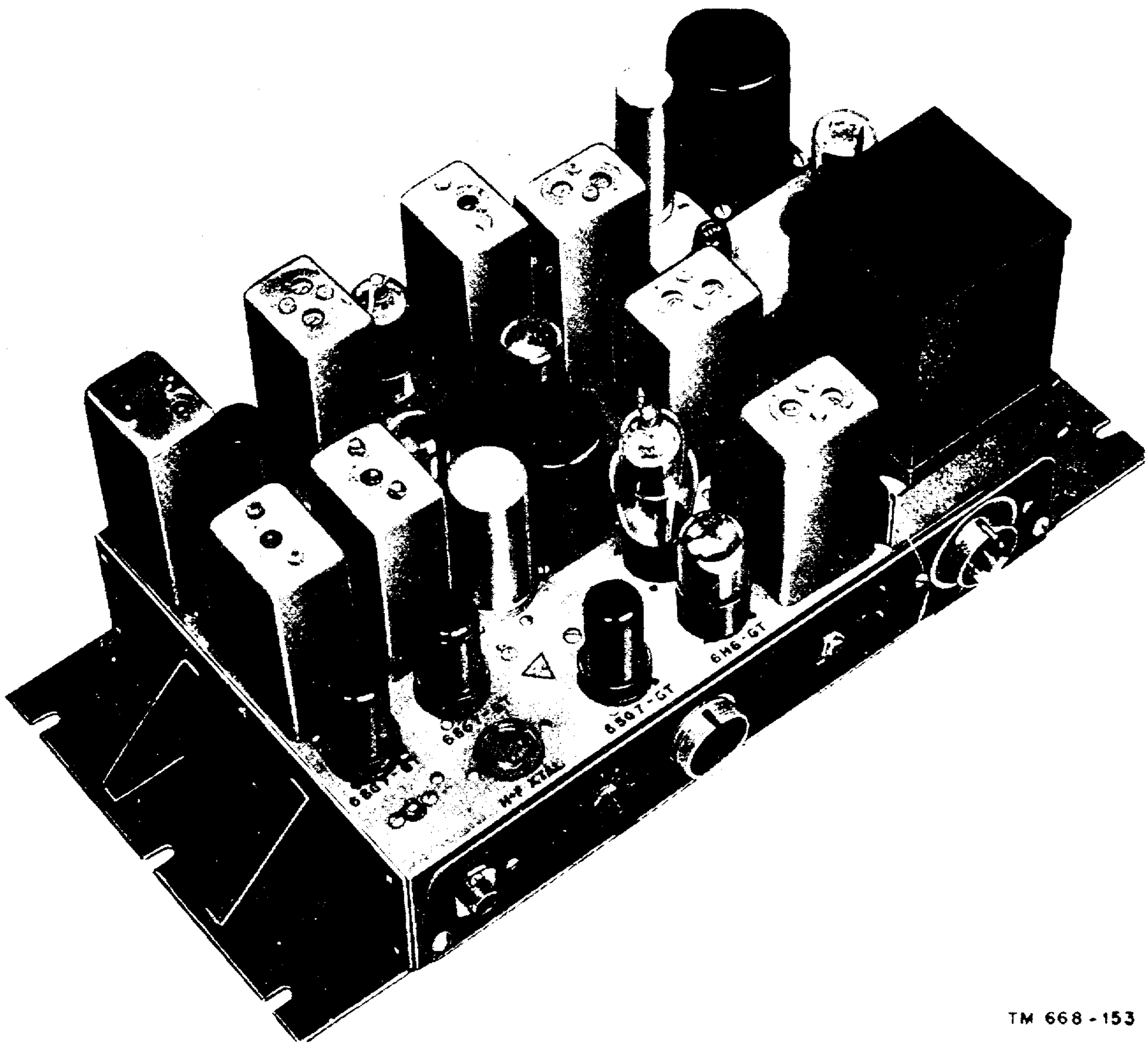


Figure 155. Typical double-conversion receiver.

talkie is available in three frequency ranges, 20 to 27.9, 27 to 38.9, and 38 to 54.9 megacycles, and the power output of the transmitter can be .9 watt, 1 watt, or 1.2 watts, depending on the model used. A single calibrated dial continuously tunes both the transmitter and the receiver. Direct f-m is used, and a built-in calibrator provides calibration points throughout the frequency range. The range varies from 3 to 12 miles. The transmitter consists of an electron-coupled oscillator and a nonlinear coil modulator plus an afc circuit. The afc circuit is controlled by a receiver interlock system which compares the frequency of the transmitter oscillator to that of the receiver local oscillator. Because the transmitter is not very powerful, the receiver accordingly is designed with high sensitivity. A .5-microvolt signal produces 2.5 milliwatts of output. The circuit consists of two r-f amplifiers, a mixer and a separate local oscillator producing an i-f of 4.3 mc, five i-f amplifiers, a phase discriminator using crystals as rectifiers, and a single stage of audio amplification. A squelch circuit and an i-f calibrator also are provided.

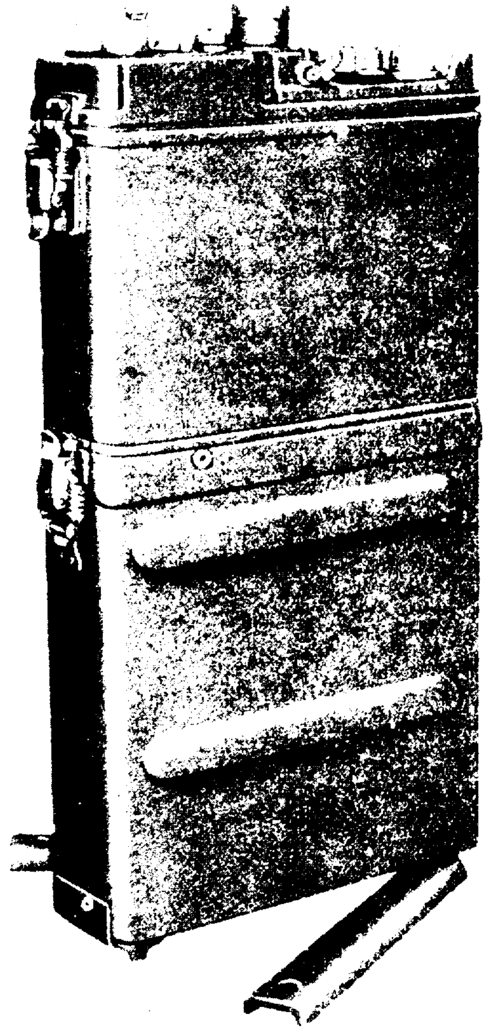


Figure 156. F-M walkie-talkie.

Section XI. ALINEMENT

93. General

a. All f-m receivers contain a large number of tuned circuits which must be adjusted correctly if the circuit is to function properly. Vibration, humidity, or temperature may cause the resonant circuits to drift off frequency in the field. Also, when a critical part is replaced, the associated resonant circuit may change frequency. The process of adjusting the tuned circuits of the receiver for optimum performance is called *alinement.*

b. The r-f, i-f, and detector circuits of any superheterodyne receiver must be alined to their proper frequencies. It is possible to accomplish a rough alinement by applying a signal at the antenna terminals and peaking the tuned circuits of the receiver until maximum output appears from the audio-frequency stages. This method is very unsatisfactory; however, in an emergency, with no servicing equipment available, the performance of the receiver can be restored to something approaching normal operation by this procedure.

c. The basic alinement procedure for any particular unit is given in detail in the manual which describes it. *Always refer to this manual for specific information.* This text indicates general principles applicable to most f-m receivers, and therefore must not be regarded as service information for any particular unit. The basic alinement procedure is to begin with the f-m detector circuit and tune the i-f amplifiers or limiters one by one, working toward the mixer, finally dealing with the r-f circuits.

94. General Alinement Methods

The two systems of alinement in common use are the *meter* method and the *visual alinement* method. The meter method uses a signal generator, which covers both the i-f and the entire r-f range tuned by the receiver, and a vacuum-tube, or high-resistance voltmeter (20,000 ohms per volt or better). The visual-alinement method uses an f-m signal generator that covers the r-f and i-f ranges, an oscilloscope, and sometimes a stable c-w signal generator. In general, an f-m receiver can be alined more quickly and easily and with far more accuracy by the visual method than by the meter method. The disadvantage of the visual method lies in the complexity of the test equipment and the difficulty of obtaining all of the necessary testing devices in the field. The meter method of alinement uses the variations of d-c voltage that take place in different parts of the receiver circuit when the tuning is changed. A steady carrier is applied from the signal generator to the circuit under test, and changes in voltage with tuning are observed on the meter. The visual method of alinement traces the actual response curves of the circuit under test on the screen of a cathode-ray oscilloscope. The tuned circuits are adjusted until the curves have the required amplitude and shape. The f-m signal generator sweeps through the frequency band covered by the stage under test, and the oscilloscope traces a curve that corresponds to the output of that stage in synchronism with the f-m generator.

95. Meter Alinement of Detectors

a. Discriminator.

(1) For meter alinement of the discriminator, a high-resistance voltmeter is needed as well as a signal generator capable of producing a stable i-f signal. When the secondary of the discriminator circuit shown in figure 157 is tuned to resonance at the desired center frequency, equal and opposite voltages are developed across $R1$ and $R2$. Consequently, there is no d-c voltage between points A and B. The tuning of the primary circuit of the discriminator affects the linearity of response. Therefore, the primary must be tuned for optimum linearity and the secondary must be tuned for correct center frequency.

(2) Connect the signal generator to the grid of the limiter tube which immediately precedes the discriminator. Connect the vacuum-tube or high-resistance voltmeter across either $R1$ or $R2$ ($M1$ in the figure). Adjust the tuning of the primary of the discriminator transformer for maximum deflection of the meter. The voltage developed across the resistor can be either positive or negative. When an ordinary meter reads backwards on the first reading, the leads must be reversed. When the meter reads maxi-

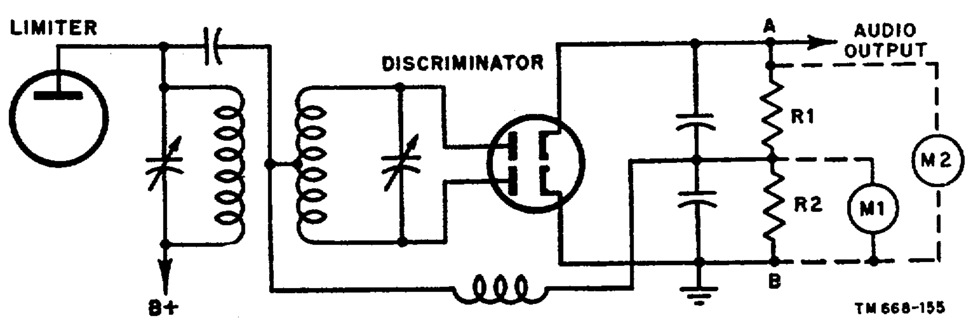


Figure 157. Discriminator alinement connections.

mum, the primary of the discriminator transformer is alined properly. When a meter with the mid-scale point at the zero-voltage mark is used, the polarity of connection is not important. Merely adjust for maximum deflection in either direction.

(3) Without changing the signal generator in any way, connect the meter across the entire load circuit (*M*2 in the figure). Since no voltage is developed across *A* and *B* when the tuning is correct, the meter reads zero when the secondary is adjusted properly. A zero-center meter is useful for this type of adjustment, since a positive and a negative voltage can be developed across the load resistors as the secondary is tuned, and the ordinary meter with zero at the extreme left of the scale is less convenient. Rock the tuning control of the secondary through zero several times, reducing the swing of voltage on either side each time, to insure accurate adjustment.

(4) The linearity of alinement can be checked easily with the meter. If the receiver must handle a deviation of 50 kc, shift the frequency of the signal generator 50 kc higher and then 50 kc lower than the carrier frequency. The meter across *A* and *B* should read the same total amount at either point, disregarding the change in polarity. With a zero-center meter, this adjustment can be observed easily as an equal deflection on either side of center for both frequencies. With an ordinary meter, the leads have to be reversed for one reading.

b. Ratio-Detector Meter Alinement.

(1) For meter alinement of the ratio detector, shown in figure 158, connect the signal generator to the grid of the last i-f tube and set it to the intermediate frequency desired. Connect the high-resistance voltmeter across the load resistor, *R*1, or the large load capacitor, *C* (*M*1 in the figure). Adjust the primary of the transformer for maximum deflection of the meter. This shows that maximum current is flowing through the load resistor and the capacitor.

(2) A symmetrical output circuit is necessary to tune the secondary. In some circuits, such symmetry already exists. When alining the circuit of figure 158, however, connect two 100,000-ohm resistors across the load capacitor, *C*. These resistors must be closely matched so that circuit symmetry is preserved. Connect the meter between the junction of the two resistors and the audio output end of the tertiary winding of the transformer. Tune the secondary for zero deflection of the meter. This shows that the secondary is exactly centered in the i-f pass band. Remove any resistors that have been

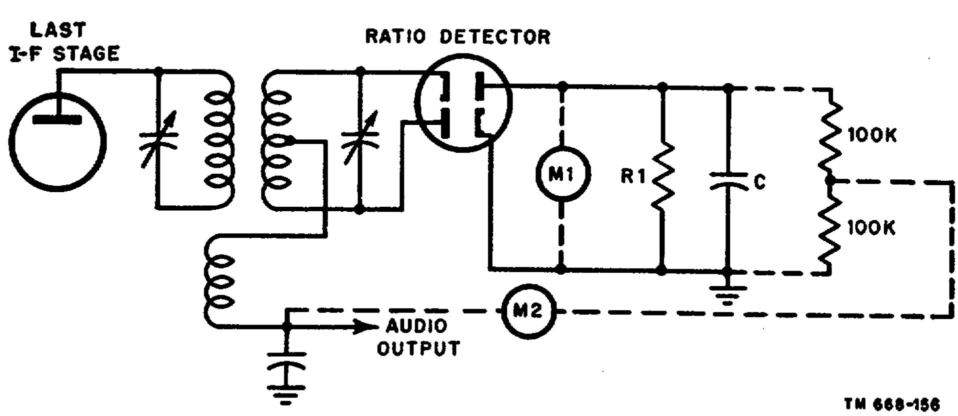


Figure 158. Ratio-detector alinement connections.

added. Linearity can be checked as in the discriminator by observing the swing of the meter on either side of center with a variation of the frequency of the signal generator.

(3) An alternative method for alinement of a ratio detector requires only an amplitude-modulated signal generator. Connect the signal generator to the grid of the last i-f stage and completely detune the secondary of the ratio-detector transformer by turning its slug or capacitor completely out. This unbalances the detector circuit and makes it sensitive to amplitude modulation as well as frequency modulation. Adjust the signal generator for about 30-percent modulation, and tune the primary of the detector transformer for *maximum* audio output. If an output meter is available, this adjustment can be made much more accurately by observing the meter deflection when it is connected across the audio output terminals. Be sure to advance the audio gain control so that a sufficient amount of signal can reach the meter, loudspeaker, or earphones.

(4) After the primary has been alined, adjust the secondary of the ratio-detector transformer for *minimum* audio output. This corresponds to maximum a-m rejection. This adjustment is critical and must be made slowly and carefully. Rock the tuning control back and forth to make sure that the minimum point is reached accurately.

c. Alinement of Locked-Oscillator Detectors.

(1) The alinement of a locked oscillator detector requires an a-m signal generator and an output meter. Connect the output meter across the audio output terminals. Turn the audio gain control to maximum. Ground the grid 1 of the detector tube; this stops oscillations and the tube functions as an a-m detector. Connect the signal generator to the grid of the last i-f tube and peak the detector transformer for maximum audio output. Then connect the generator to the second i-f tube grid and peak the next i-f transformer, continuing this process until the grid of the mixer is reached.

(2) Turn off the modulation in the signal generator and remove the ground from the grid of the detector tube. Short the quadrature circuit by connecting a jumper across the plate coil. The oscillator circuit continues to function. Adjust the oscillator trimmer capacitor until a heterodyne beat note is heard in the earphones or loudspeaker. This is the oscillator beating against the unmodulated carrier from the signal generator. Adjust the oscillator trimmer for zero beat, watching the output meter for an exact indication. The basic frequency of the oscillator now coincides with the frequency of the i-f signal generator.

(3) Remove the short from the quadrature circuit. Reduce the output from the signal generator to the point where the oscillator in the detector circuit is unable to lock in. Adjust the tuning of the quadrature coil so that the zero beat note is restored. This completes the alinement of the detector.

d. Alinement of Gated-Beam-Tube Detector.

(1) Since the gated-beam tube combines the functions of a limiter and a discriminator in one circuit, the alinement of the circuit is divided into two parts. Connect an *a-m* signal generator with a low percentage of modulation to the grid of the last i-f amplifier and an output meter to the audio output terminals. Turn the audio gain up fully. Short the quadrature circuit grid to ground, and decrease the output from the signal generator until a variation in the signal-generator output control produces a variation in the audio output. This setting insures an amount of i-f signal that is below the threshold of limiting. Tune the secondary and the primary of the detector input transformer for maximum audio output. As the amplitude of the signal increases with alinement, limiting begins to take place at the grid, and it is

difficult to determine sharp tuning points. When this difficulty occurs, reduce the output of the signal generator further, and repeat the peaking of the transformer.

(2) Increase the output of the signal generator to the point where limiting occurs as indicated by no further increase in audio output with increased signal-generator output. Increase the percentage of modulation and adjust the variable cathode resistor for minimum audio output. Repeat this threshold adjustment until further small changes in the cathode resistor setting do not affect the point at which the threshold of limiting begins. This completes the alinement of the limiter section.

(3) Connect an f-m signal generator with the required deviation to the grid of the last i-f amplifier. Remove the short from the quadrature circuit and tune the quadrature circuit for maximum audio output. This completes the alinement of the gated-beam tube.

96. Meter Alinement of I-F Stages

a. The meter alinement of the i-f stages in a receiver measures the voltage developed at the detector circuit as the i-f amplifiers are tuned. In the ratio detector, the meter is connected across the load capacitor, and in the gated-beam and locked-oscillator circuits, an audio output meter is used when the detector is altered so that it responds to a-m. When alining a receiver that uses a discriminator-limiter detector, connect the meter across the limiter grid resistor. If two resistors are used in series, connect the meter between the junction of the two and ground. As the signal to the limiter increases, the grid current increases, and consequently the voltage across the grid resistor rises.

b. Always work from the detector or the limiter toward the mixer. In the limiter-discriminator circuit, aline the discriminator first. If a dual limiter is used with tunable coupling between the two stages, aline the first limiter by connecting the meter across the grid resistor of the second limiter. Connect the signal generator to the grid of the first limiter. Then tune the interstage coupling circuit for maximum reading of the meter. Connect the meter to the grid resistor of the first limiter and proceed with the i-f alinement.

c. When all of the i-f amplifiers are known to be single-peaked—that is, when none of them are overcoupled—tune each secondary and primary, working toward the mixer from the first-limiter grid or from the grid of the last i-f stage, depending on the type of detector. Adjust each tuning control for maximum deflection of the meter. The signal generator must be unmodulated when the discriminator and ratio detector are alined, and amplitude-modulated for other detectors.

d. If one of the i-f amplifiers is overcoupled, the alinement procedure for that stage is slightly different from the procedure above. Using short leads, connect a low value of resistance (200 to 500 ohms) across the secondary of the overcoupled transformer. This serves to reduce the *Q* of the transformer, and the reduced *Q* suppresses the double-humped characteristic of the overcoupled circuit, with the result that there is a single broad resonant peak in the circuit. Observing the output meter carefully, tune to the exact center of this broad peak. The overcoupled circuit usually is the second i-f transformer in a limiter-discriminator type receiver. Disconnect the loading resistor when the stage has been alined.

97. Meter Alinement of R-F, Mixer, and Oscillator Stages

a. The alinement of r-f and mixer stages requires an accurately calibrated r-f signal source that can be tuned to frequencies in the low, high, and center portions of the receiver. The mixer and the oscillator must track over the desired range if the receiver is continuously tuned, and tracking adjustments may vary widely with the type of receiver. Consult the appropriate equipment manual for a particular piece of equipment.

b. The mixer trimmer and the oscillator trimmer are adjusted at the high frequency end of the tuning range for optimum gain and calibration. Set the receiver tuning control at the

designated alinement frequency. Connect a signal generator to the grid of the r-f amplifier. Set its frequency to the highest calibration point called for in the equipment manual. Vary the oscillator trimmer until the signal is heard in the output circuit, or until the meter in the detector circuit indicates maximum deflection. The generator is unmodulated for ratio and discriminator detectors, and amplitude-modulated for the others. Peak the mixer grid-circuit trimmer for maximum output. Tune the receiver to the designated low-frequency alinement point. Tune the signal generator to the low end of the frequency range and set it at the low-frequency calibration point. Adjust the oscillator padder until the signal is a maximum at the detector. Then return the signal generator and receiver to the high-frequency calibration point. Retune the oscillator trimmer, if necessary, to bring the dial calibration of the receiver into correspondence with the frequency of the generator. Repeat these steps until both the low and the high frequencies are on calibration without the need for touching either the oscillator trimmer or the padder. Set the signal generator to a point midway in the frequency range, and check the calibration. When these steps have been carried out carefully, the calibration should be correct.

c. Alinement of the r-f stage is done best with a noise generator, as described in the section on r-f amplifiers. Lacking a noise generator, a fair approximation is offered by the shorting technique described in that section. For the first rough initial alinement, antenna noise or a weak external signal supplied by the leakage from the signal generator or any other source will do. Peak the trimmer for maximum signal at the high end of the frequency range, and apply the noise-generator technique at the highest frequency to obtain the maximum signal-to-noise ratio.

98. Visual Alinement

a. Principles.

(1) In visual alinement, a cathode-ray oscilloscope and an f-m signal generator are used in place of the meter and the a-m signal generator. The actual response curves of the i-f and detector circuits are traced out on the scope by this method, whereas in the meter method of alinement only the maximum response points are known. Therefore, the visual method of alinement provides a more accurate means of adjustment and permits a very rapid grasp of exactly what is happening, since the actual behavior of the circuits with an f-m signal is visible.

(2) The oscilloscope is a device that responds to voltages; therefore, it always is connected in parallel with the circuit under test. The oscilloscope can be used to display currents, but first they must be passed through a resistor and then the voltage developed across the resistor is measured. The f-m signal generator is a standard signal generator with internal means for frequency modulation by either a sine wave or a triangular wave. The internal modulating signal is applied to the plates of the oscilloscope to provide a sweep voltage. For this reason, the device is called a *sweep* signal generator.

(3) Also useful in conjunction with visual alinement methods is an accurately calibrated signal generator (marker generator) which produces c-w signals to serve as directly visible frequency check points on the crt (cathode-ray tube) screen. Where a large number of receivers of a particular type are to be alined, crystal-controlled marker generators often are used for various critical frequencies, like the center of the i-f pass band and its outer limits.

b. Alinement of I-F and Limiter Stages.

(1) The input terminals of the scope provide for spot deflection both horizontally and vertically. For i-f alinement, the vertical-deflection terminals are connected across the limiter-grid resistor. This produces a voltage that is proportional to the amplitude of the signal input to that stage. The horizontal-deflection plates are connected

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to the sweep voltages from the f-m signal generator. Connect the r-f output of the sweep generator to the grid of the last i-f stage. Adjust the frequency modulation for a deviation a few times greater than that of the rated deviation limits of the receiver. As the frequency swings through the i-f pass band, alining the tuned circuits changes the position and amplitude of the signal that appears at the limiter grid. Consequently, the variation of amplitude appears as changing vertical deflection on the screen of the cathode-ray tube. At the same time, the horizontal deflection changes in direct relationship with the sweep voltage. Since this also is the frequency modulation signal for the generator, the horizontal axis is effectively a frequency axis. Therefore, the picture displayed on the face of the tube when the signal is passed through an i-f amplifier is that of the familiar bell-shaped tuned-circuit response.

(2) As the tuning of the stage is varied, the response curve moves horizontally on the screen, in addition to changing in amplitude. Connect the marker generator to the grid of the last i-f stage. This causes a small wiggle to appear in the curve as the signal and marker generators beat against one another. If the marker generator is set to the intermediate frequency, the tuning of the circuits is adjusted to center the bell-shaped curve at the marker pip, as in A of figure 159. Similarly, two marker generators can be used to mark the limits of the i-f pass band, with a resultant trace on the screen that resembles that in B.

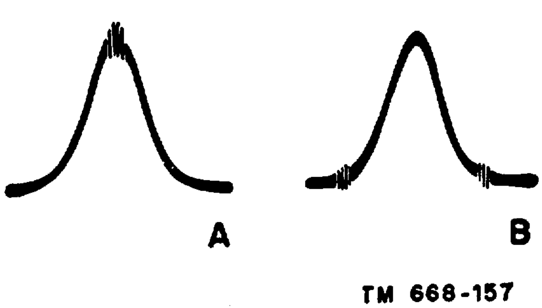

Figure 159. Use of marker frequencies in i-f alinement.

(3) Each of the i-f stages can be alined, connecting the sweep and marker generators to the grids of the preceding stages in turn. When a double-tuned overcoupled i-f circuit is used, the double-humped characteristic appears directly on the screen and can be centered at the marker frequency. There is no need to use any loading resistance. As additional stages are tuned, a broad, flat-topped characteristic can be developed that provides the least distortion and best off-channel attenuation.

c. Alinement of Discriminators.

(1) Connect the vertical input of the oscilloscope to point A and ground in the discriminator of figure 157. Connect the f-m signal generator to the grid of the limiter immediately preceding the discriminator. Connect the horizontal-sweep voltage of the f-m generator to the horizontal input of the oscilloscope. As the signal on the horizontal plates varies, the spot on the screen traces out a line that corresponds to the amplitude of the modulating signal, and therefore to frequency. As the frequency changes, the instantaneous voltage developed across point *A* changes, and the spot is deflected either upward or downward, depending on its polarity and amplitude. When the discriminator is alined properly, the characteristic *S* curve is obtained, lying at an angle to the horizontal, and crossing the zero-voltage spot position.

(2) A marker generator connected to the limiter grid and tuned to the center frequency permits the development of the *S* curve in A of figure 160. When the marker generator is set at either the upper or the lower frequencies limit, the indications in B and C appear. As the secondary of the discriminator is tuned, the curve moves up or down, as in D. When the primary is

mistuned, the *S* curve is ragged, as in E. With correct tuning of the primary and secondary, the curve looks like that in F. First tune the primary for a smooth linear curve within the desired range, and then tune the secondary so that the line is centered on the marker frequency or on the center of the scope.

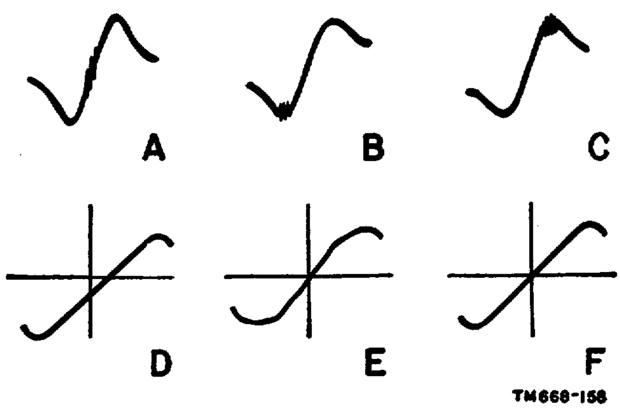

Figure 160. Use of marker frequencies in discriminator alinement.

d. Alinement of Ratio Detector.

(1) When visually alining receivers with ratio detectors, connect the vertical plates of the scope to the same point used in the meter method (fig. 158) or from the audio lead to ground. Connect the f-m signal generator to the grid of the last i-f tube.

(2) Detune the secondary trimmer. Peak the primary trimmer for a maximum single and symmetrical i-f curve. Aline the remainder of the i-f's, moving back a stage at a time toward the mixer. Then adjust the secondary of the ratio-detector transformer to give the S-shaped characteristic curve. A marker generator is helpful in setting the exact center.

(3) A second method of alinement starts similarly: Connect the scope and generator as above. Tune the primary for the single-peaked curve. Then tune the secondary for the *S* curve. Move the signal generator a step at a time toward the mixer, tuning each stage to obtain the largest symmetrical and linear *S* curve.

(4) Another method requires the opening of one lead of the load capacitor. Then the circuit operates as an a-m detector. The terminals of the scope must be connected across the load resistor. Adjust all of the i-f transformers for a symmetrical curve. Reconnect the capacitor, connect the scope to the meter position, or across the audio output to ground. The scope now shows the *S* curve. Adjust the secondary of the ratio-detector transformer for proper centering.

e. Alinement of Locked Oscillator.

(1) Connect the vertical input of the scope to the junction of the plate load resistor and the quadrature coil and ground. To aline the i-f stages, ground the oscillator grid. Connect the f-m signal generator to the grid of the last i-f tube and adjust the tuned circuits for the familiar single, peaked symmetrical response. Continue with the following i-f amplifiers, working toward the mixer.

(2) After the i-f amplifiers are alined, remove the short from the oscillator grid, tune the generator to the i-f center, turn off the f-m, and short the quadrature network. The signal generator and scope remain where they were. Adjust the oscillator trimmer for zero beat, as indicated by minimum signal on the oscilloscope. Remove the short from the quadrature circuit, turn on the f-m in the generator, and set the deviation for the rating of the receiver. Adjust the quadrature coil for linear response on the oscilloscope screen (fig. 161). In A, the quadrature circuit is misalined. In B, the alinement is correct. In C, both the quadrature circuit and the last i-f transformer secondary circuits are misalined. Marker generators are not needed for the center frequency of the

locked oscillator. Frequently, it is easier to use meter methods of alinement for the r-f stages and visual alinement for the i-f and detector circuits. This requires the use of a considerable amount of equipment, but saves time when a large number of receivers of the same type must be alined.

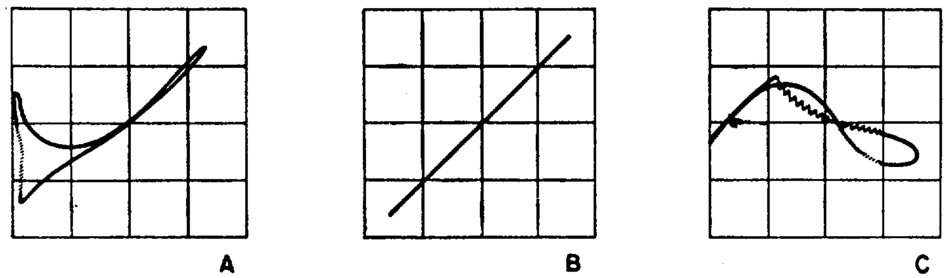

Figure 161. Scope patterns for locked-oscillator detector.

Section XII. SUMMARY AND REVIEW QUESTIONS

99. Summary

a. F-m receivers generally are of the superheterodyne type.

b. The f-m signal voltage in a superheterodyne is amplified and combined with the voltage from a local oscillator in a mixer. The difference-frequency output is fed to an i-f amplifier.

c. The output of the i-f amplifier is converted into audio voltage in an f-m detector. The audio signal is amplified and applied to speakers, headphones, or other audio devices. In the absence of signal, a squelch circuit can be used to prevent circuit noise from reaching the output.

d. For greater selectivity, double-conversion superheterodynes are used. The output of the first i-f amplifier is converted to lower second i-f.

e. The r-f amplifier receives the lowest level signal in the receiver from the antenna and amplifies it before it reaches the mixer. The critical features in the performance of the r-f amplifier are image rejection, noise figure, and attenuation of radiation from the local oscillator.

f. The noise figure of the r-f amplifier depends on its bandwidth, the type of circuit, and the tube used. The noise performance of the receiver depends almost entirely on the noise figure of the r-f amplifier, if its gain is high. The image rejection depends on the selectivity of all tuned circuits encountered by the signal before reaching the grid of the mixer.

g. The mixer stage operates as a low-level modulator, with the local oscillator acting as the carrier and the signal as the modulation voltage.

h. The voltage gain of a mixer depends on its conversion transconductance, which is approximately the ratio of the i-f plate current to the r-f grid voltage.

i. The local oscillator signal can be injected on the same or different tube electrode compared to that to which the signal is applied. The type of oscillator injection determines the extent of its radiation, interaction between oscillator and mixer circuits, and loading of the mixer input circuit.

j. Multigrid mixers are used frequently for the second i-f sections of double-conversion receivers.

k. Converters are oscillators and mixers combined in the same tube envelope. The electron stream of the tube serves for both sections with a large number of grids performing the different functions necessary.

l. At very-high and ultrahigh frequencies, crystal mixers can be used. These units have conversion loss instead of gain, and are seldom